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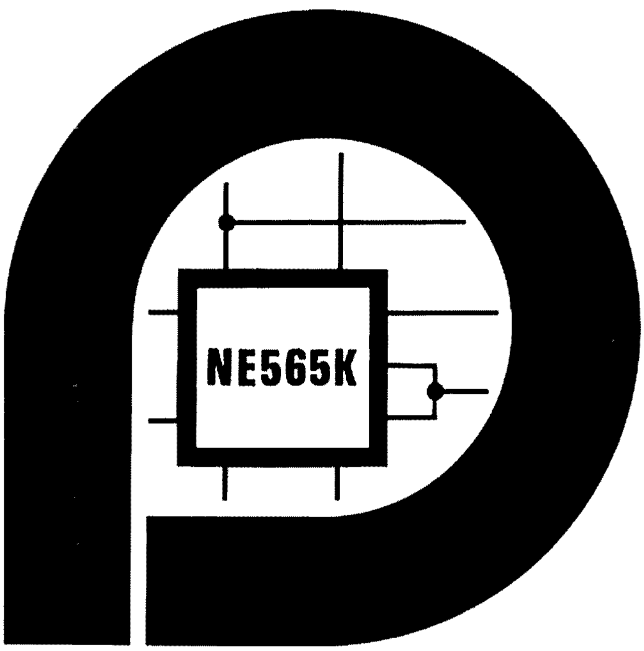

PHASE-LOCKED LOOP RTTY TERMINAL UNIT

this month

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ham radio magazine

contents

a second look

by jim fisk

A whole new field of electronics may eventually evolve from current research into semiconductor vacuum tubes. Sound like an April-fool joke? Well, interest in this area has been high since RCA successfully operated a silicon-cold-cathode tube several months ago. In this device the cathode is a silicon p-n junction which emits electrons directly into a vaccuum when the diode is forward biased.

The latest solid-state cold-cathode vacuum tubes use gallium phosphide as the cathode material. The surface of the gallium phosphide is coated with a layer of cesium or cesium oxide which allows the electrons to escape into the vacuum. Although the maximum current densities are presently quite small, and must be increased significantly before the solid-state vacuum tubes are practical for commercial use, technology moves very rapidly, and we should see practical devices of this type on the market within the next five years.

Because of another recent advance in electronic technology noisy cooling fans and blowers may shortly be giving way to a completely new cooling concept which uses, of all things, an electrostatic discharge. The heat is removed by the corona discharge from the negative ends of high-voltage probes — there is no arcing. According to the inventor, Oscar Blomgren, a red-hot heat sink at 1675° Fahrenheit can be cooled to a dark 975° (a drop of 700 degrees) in one or two seconds with a 30-kV, 200-μA electrostatic discharge.

Blomgren accidentally discovered the cooling phenomenon of high-voltage discharges while attempting to keep an acetylene flame from touching the inside of a pipe by using an electric field; he was trying to solve a burner-nozzle deterioration problem.

There is still much to be learned about the way the discharge cooling system works, but apparently the corona discharge or electric wind creates vortex columns in the air next to the heated surface. Normally, a thin layer of air clings to the surface; this acts as an insulating barrier which inhibits the rate at which adjacent cooling air can carry away the heat. With the high-voltage electric field, the cooling rate is increased by the swirling action of the vortex columns which pull in cooler air from outside the normal boundary area.

To compare electrostatic cooling with forced-air cooling, engineers set up an aluminum block and heated it to 250° F. With constant heat input, the temperature decreased at the rate of 8° per minute when the high-voltage was turned on. When a blower was used, the temperature dropped 11° per minute, but 14 watts were required for the blower as opposed to 3 watts for the electrostatic system.

In another experiment engineers heated two high-power transistor heat sinks to 500° F with blow torches. With the burners adjusted to stabilize the temperatures of both heat sinks to 500°, a 28-kV electric field applied to one heat sink reduced its temperature by 185 degrees; the other heat sink remained at 500°.

Amateurs who would like to experiment with electrostatic cooling need an adjustable high-voltage power supply, 20 to 30 kV. The unit to be cooled is connected to ground with the negative lead of the supply; the positive lead is connected to a pointed probe. No noticeable cooling takes place until the discharge current is greater than about 10 μA; this is controlled by the supply voltage and the spacing between the probe and the unit being cooled.

Jim Fisk, W1DTY
editor

phase-locked loop RTTY terminal unit

New design for a solid-state afsk demodulator and selector-magnet driver with the features most wanted by RTTY operators

Ed Webb, W4FQM, Post Office Box 17, Schaumburg, Illinois 60172

Upon acquiring a new teletype machine, I was faced with the problem of coming up with a terminal unit, or TU – the black box between the transceiver and the teleprinter. I didn't want to be bothered with unwinding toroids and selecting odd-ball capacitor values to adapt some of the TUs that were holdovers from the past decade. I wanted a TU that was totally new in its detection method. This TU was to be "state of the art" and thus different in design from current units.

I'd had some experience with the phase-locked loop (PLL or PL^2) in the telemetry field and knew it to be an excellent fm demodulator. However, the discrete circuitry was rather complex. While deliberating, I received a notice from Signetics announcing their new PL^2 in an IC. This announcement, together with subsequent attendance at their technical seminar, pointed the way for my new TU design – a phase-locked loop detector.* With the help of Art Fury, WA6JLJ, and his staff at Signetics, my project has progressed from a dream to reality – all in one year.

*The circuit is now patent pending, and the PC boards (available from WCI, Box 17, Schaumburg, Ill. 60172) are copyrighted.

the phase-locked loop

The PL^2 as an afsk detector is an entirely new approach to RTTY demodulation. No toroids or LC-tuned circuits are used. The entire TU occupies two 4-inch-square PC boards. The PL^2 detector will work with 6-dB or lower signal-plus-noise-to-noise ratios. No scope tuning provisions are needed; a pair of light-emitting diodes provide all tuning indications. Features of the PL^2 afsk demodulator are:

1. Automatic shift selection, allowing automatic copy of any shift from 150 to 1000 Hz without manual switching.

2. Afc, which provides automatic lock-in; signals drifting as much as ±500 Hz can be followed.

3. Automatic threshold corrector.

4. Antispace control.

5. Autostart, with solid-state printer motor control.

6. A 170-volt loop supply, coupled with extremely sharp selector-magnet pulses, and a constant-current selector magnet driver.

7. Operation with ssb audio passbands (low tones) or regular RTTY tones at the flip of a switch.

8. Squelch circuit, which prevents printing on noise.

characteristics

The phase-locked loop has an extremely linear voltage-controlled oscillator (vco) whose center frequency is set to the midrange of frequencies being detected. A simplified functional diagram of the Signetics NE565K IC phase-locked loop is shown in **fig. 1**. An RC network sets the resting or center frequency, F_{co}, of the vco. When a signal is applied to the phase-comparator input, its frequency and thus its phase is compared with F_{co}. If the vco frequency is different from F_{co}, an error voltage is generated at the phase-comparator output. This error voltage is then amplified, filtered, and fed to the frequency-control input of the vco. The error voltage causes the vco frequency to shift to that of the applied

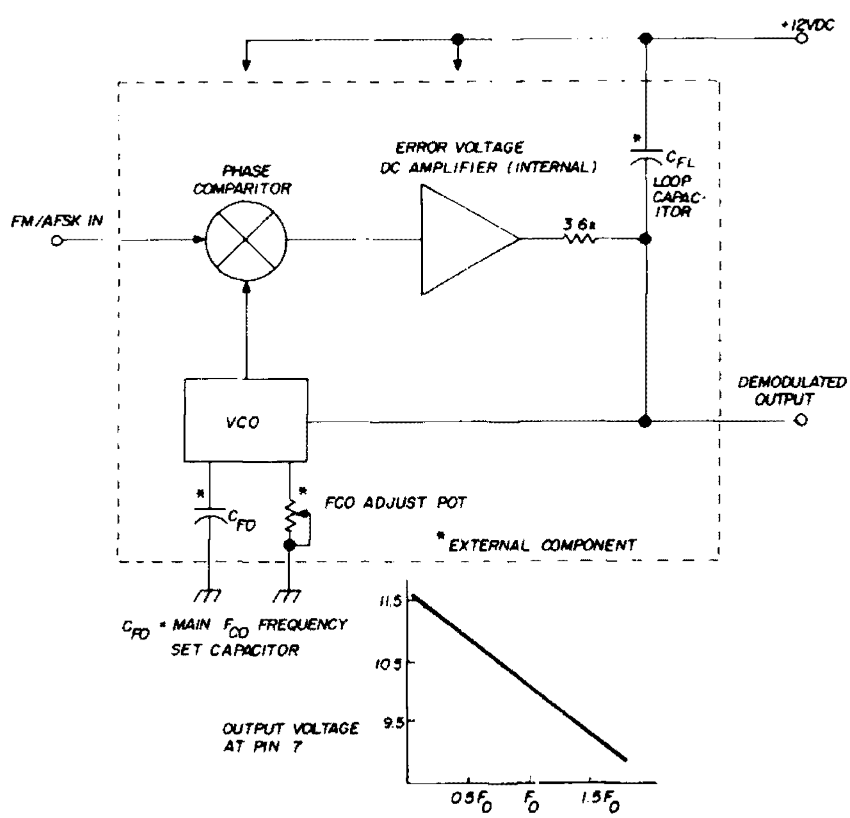


fig. 1. Oversimplified diagram of Signetics NE565K phase-locked loop. Loop 6-dB roll-off point is where X_c of C_{FO} equals 3.6 k ohms.

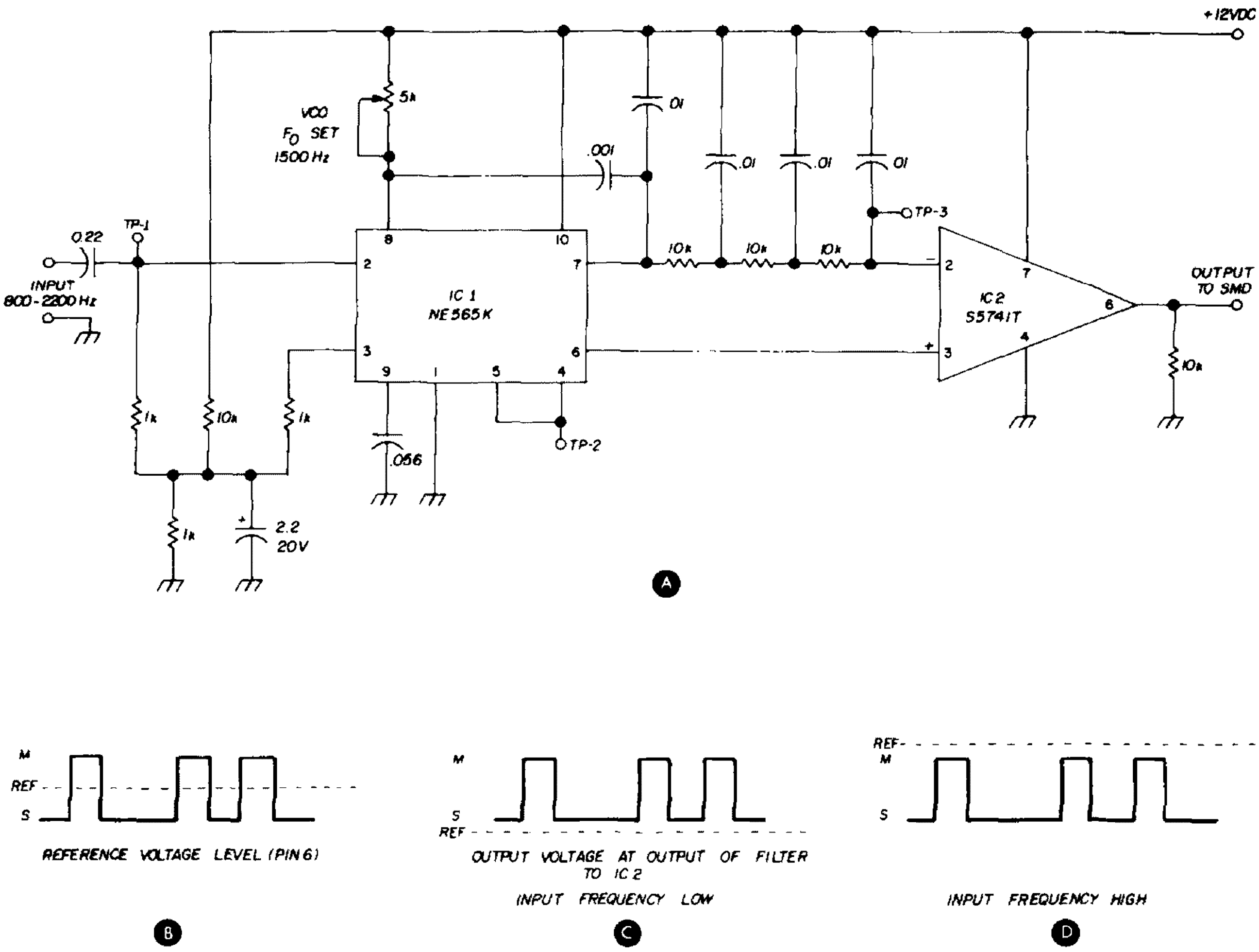


fig. 2. Trail circuit for the PL^2 afsk detector, A-waveforms at B and C represent asymmetrical tracking-comparator output due to drifting input signals.

input signal. This action nulls the error-voltage output from the phase comparator, and the loop is phase locked.

If the input frequency varies within the vco frequency range, the error voltage applied to the vco is the duplicate of the frequency modulation applied to the PL^2, and is thus the demodulated output. The vco output may also be used as it has an approximately 40-dB s+n/n ratio over the input signal, and the vco tracking-response rate may be controlled by an external low-pass filter capacitor. Thus the PL^2 may be used as an fm detector/discriminator and as a tracking filter.*

development

The recommended circuit for the Signetics NE565K phase-locked loop required a dual-voltage power supply. This was contrary to my basic premise for an afsk detector; namely, to keep it simple. More reading and experimentation showed that the PL^2 would work with a single-polarity 12-Vdc supply. The recommended S5710 voltage comparator wouldn't operate with a single-polarity supply, but I found that the S5741T would. So I used a NE565K-S5741T combination for my first detector (**fig. 2A**).

The S5741T is internally compensated and develops more than 100-dB voltage gain. The NE565K provides approximately 100-mV change at pin 7 for 200-Hz frequency change at the input. This

*The phase-locked loop isn't new. In the 1950s a modified version of the PL^2 described here was used to provide phase-coherent reference signals in a tracking station at the Kennedy Space Center. This early circuit, using tubes, occupied half of a 6-foot relay rack. Two technicians were assigned full time during range operations merely to check for balanced sets of tubes! editor.

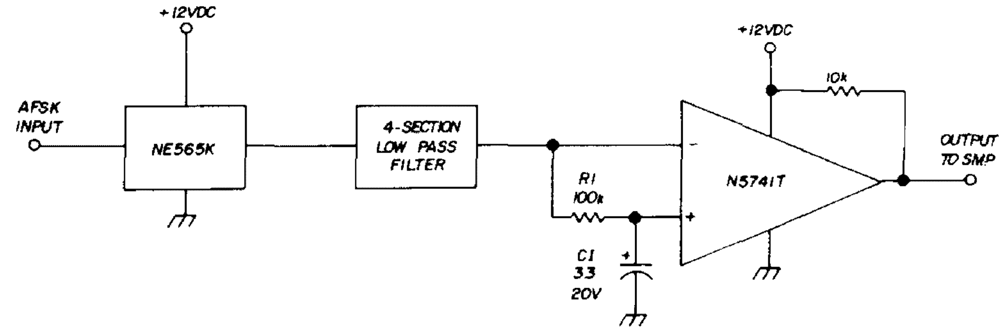


fig. 3. Integrator circuit for obtaining stable reference voltage at input to tracking comparator.

detector followed RTTY signals from 170-Hz narrow shift to 850-Hz wide shift, and even followed commercial 425-Hz-shift stations – all without switching or adjustments!

A three-section RC filter eliminated the vco carrier frequency on which the demodulated information appeared at pin 7. The one drawback to this circuit was that receiver tuning had to be just right to keep the PL^2 output swinging symmetrically on either side of the reference voltage at pin 6. If the receiver or received signal drifted, then the output from the comparator became asymmetrical, and the printer either ran open or refused to print (figs. **2B** and **2C**).

tracking problems

The problem was clear but baffling. The reference voltage must shift in proportion to the amount of tuning error so that the comparator output remained symmetrical. If the demodulated output voltage swing could be averaged, then this voltage could be used as a reference voltage, which would move with the demodulated output-voltage swing. But this had to be done without disturbing the PL^2 output.

My approach was to integrate the total swing, thus obtaining an average, or midpoint, of the demodulated-voltage swing that could be used as a reference (**fig. 3**). This reference voltage would then shift with the voltage swing of the PL^2 output if the applied frequencies were (a) offset from the vco center frequency, or (b) different because of frequency drift in the receiving equipment.

Since the shortest RTTY pulse at 60 wpm is 22 ms, I chose integrator time-constant values that provided 15 times the shortest pulse, or 0.33 second, so that the integrator followed the average swing of about 3 RTTY characters at 60 wpm. Resistor R_1 (fig. 3) unbalances the input to the N5741T tracking comparator so its output remains in mark with no signal applied to the PL^2; thus the machine does not run open. The output voltage from pin 7 of the phase-locked loop was about 10.6 volts with no input.

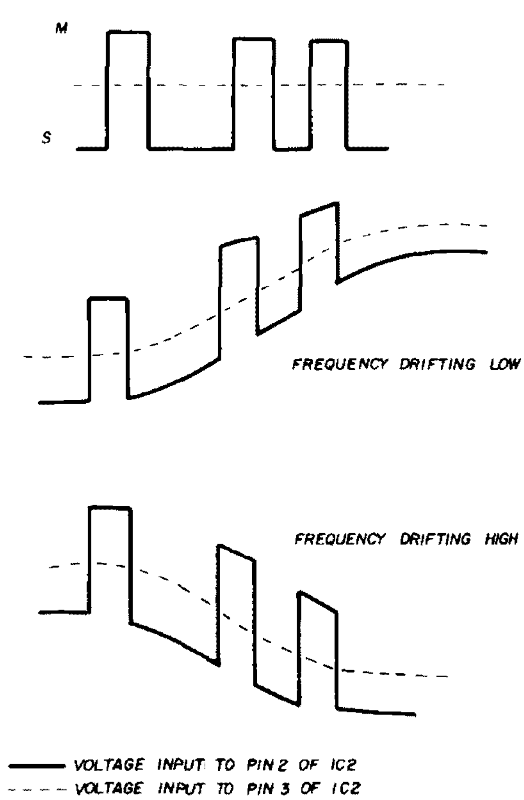


fig. 4. Dashed line represents reference point of input to tracking comparator as a result of the low-pass filter. Note that reference point is always at the midpoint of the demodulated voltage swing regardless of signal amplitude, which is proportional to frequency shift.

detector performance

The circuit of fig. 3 worked as well as I had expected. Even slight mistuning of afsk frequencies had no effect on printer operation so long as these frequencies fell within the lock-in range of the PL^2. This range was about ±50 percent of the vco center frequency. The low-pass filter was increased to 4 sections to obtain another 6 dB of vco carrier suppression and slightly improved performance. I now had a workable detector that had a form of afc and automatic shift selection – the reference point would always be at the midpoint of the afsk input demodulated swing regardless of the demodulated swing amplitude, which is proportional to the input-frequency shift. (See fig. 4.)

The differential input of the PL^2 was terminated to present a 600-ohm load at the input. This termination was ac coupled to ground through a capacitor, as the inputs require positive bias voltage for normal operation. The PL^2 has very high input sensitivity and will operate on signals in the order of 100 mV p-p or less. A simple diode limiter was placed in the input to avoid overdriving the PL^2.

No predetection filtering was used because I wanted to retain the maximum bandwidth of the PL^2. Thus the only predetection filtering or input-bandwidth restrictions are those of the receiver rf-af passband and a single-section RC filter in the vco control-voltage circuit; thus the PL^2 operates as a second-order loop. Its response is limited to 450 Hz by the 0.1-μF capacitor between pin 7 and $+V_{cc}$ (fig. 2).

squelch circuit

Much on-the-air testing followed. The PL^2 and tracking comparator worked extremely well. However, one problem continued to annoy me. When tuning from one RTTY station to another, or when an RTTY station went off the air, random noise caused the printer to print random characters. This was because of the tracking comparator high gain and the wide bandwidth of the PL^2 and its RC filter.

A method was needed to put the printer in a nonprint or mark-hold mode in the absence of a valid RTTY signal, or in the presence of random noise. A simple noise squelch would do the job but would require another stage that could be squelched. I selected a type TAA-560 Schmitt trigger by Amperex for this circuit (fig. 5). The TAA-560 requires a trigger voltage of only 1.5 volts. The tracking comparator output is of the order of 10 V p-p (in the space condition), so input to the Schmitt trigger is controlled by a voltage divider.

The PL^2 noise voltage is of insufficient amplitude for use as a squelch voltage; however, the tracking-comparator output is high enough for this application. (Also, I didn't want the 22-ms TTY pulses to activate the squelch circuit.) I sampled the tracking comparator output with a differentiating circuit. This network had little response to the comparatively long RTTY pulses, which have approximately a 25-Hz rate. The differentiating network responded to pulses with a frequency of 200 Hz or higher. Its design frequency is approximately 200 Hz at the 6 dB point, and it works like a simple RC high-pass filter.

The pulses from the differentiating network are coupled to a voltage doubler circuit. The rectifier output voltage is applied to a filter capacitor with a bleeder resistor in the form of a pot across it. The combination of the 10 μF filter shunted by the 100-k pot provided a discharge time constant of about 1 sec. The charge time is limited only by the circuit impedance and is about 10 ms. The output arm of the squelch pot was also fed to the Schmitt trigger input through a 100-k isolation resistor.

When the detector is fed raw audio noise from the receiver, the squelch pot is adjusted until the machine stops chattering on noise. The negative voltage from the squelch rectifier keeps the Schmitt trigger biased off. The Schmitt output will remain in the mark condition with random noise at the PL^2 input, or when

the RTTY signal fades into the noise to the point where the PL² unlocks. In the absence of an input to the PL², integrator resistor, R11, at the tracking comparator input unbalances the static positive volt-

RC low-pass ladder filter. The filter output is fed to the Schmitt trigger, which functions as a pulse regenerator. The Schmitt trigger output pulse rise time is 10 μs.

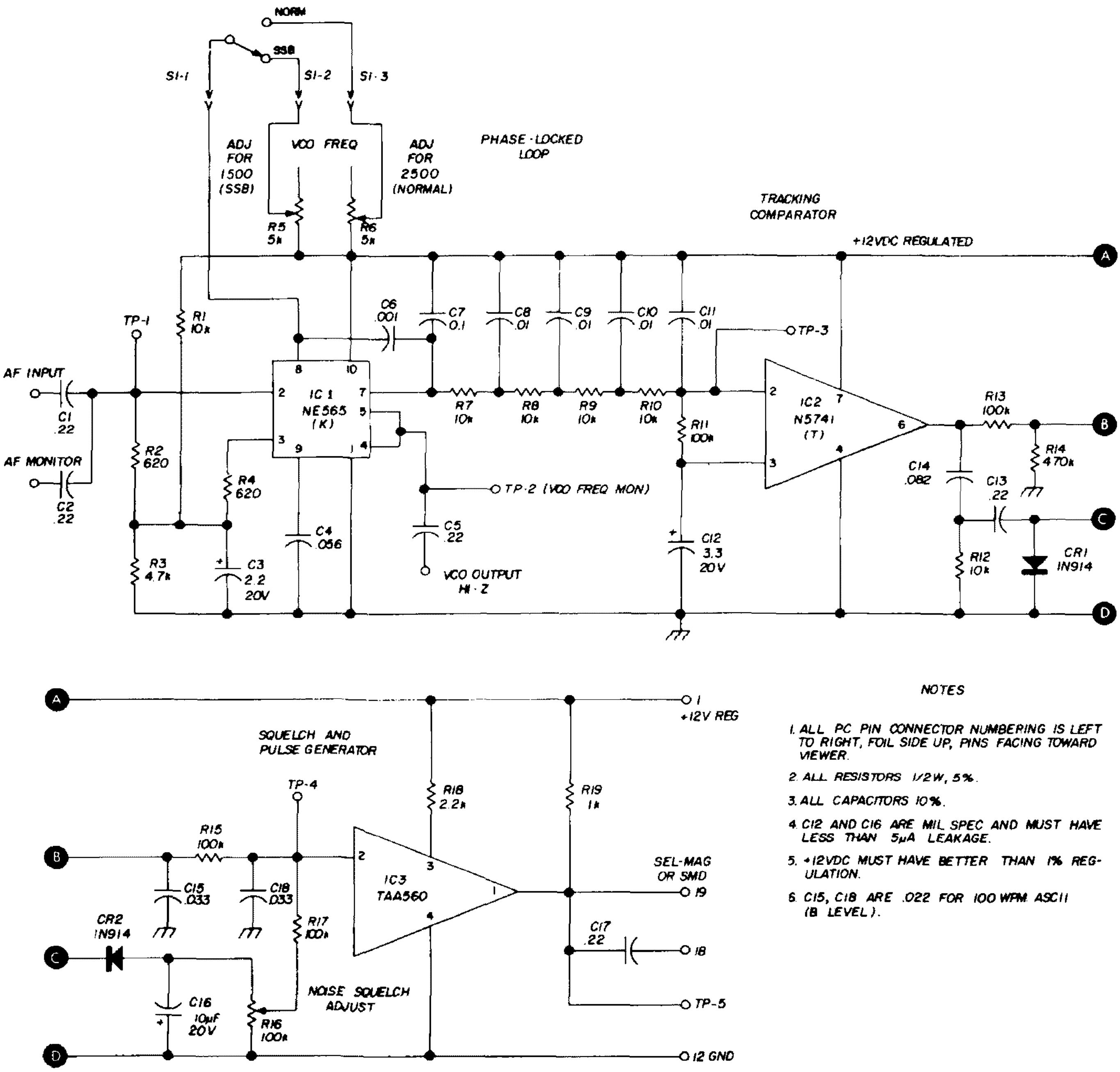


fig. 5. Schematic of the afsk detector incorporating the refinements described in the text. Circuit is Patent pending, 1971, by WCI.

age at the differential input; thus the tracking comparator output remains in mark.

post detection filtering

The tracking comparator output is fed through a voltage divider to a two-section

If it's desired to pass a square-wave pulse through a network without distorting the pulse waveshape, the network should pass a frequency at least ten times (preferably 20 times) higher than the *pulse frequency. However, when maximum* signal-to-noise ratio is desired, mini-

mum bandwidth must be used. A minimum-bandwidth filter, unfortunately, also distorts an applied square wave; so a square-wave signal from a narrow-band filter must be reshaped. Hence the TAA-560 IC in **fig. 5.**

When the pulse is regenerated great care must be taken not to induce pulse-width distortion. The only place that a non-square-wave pulse may be sampled (for regenerating purposes) is at the 50% maximum voltage point, where the pulse rise-time delay is exactly equal to the fall-time delay. In this way the pulse is delayed because of the narrow filter bandpass, but its width doesn't change during regeneration or shaping. The regenerator output produces the exact duplicate (possibly better) of the original pulse before entering the narrow-band-pass filter; or, in this case, a low-pass filter.

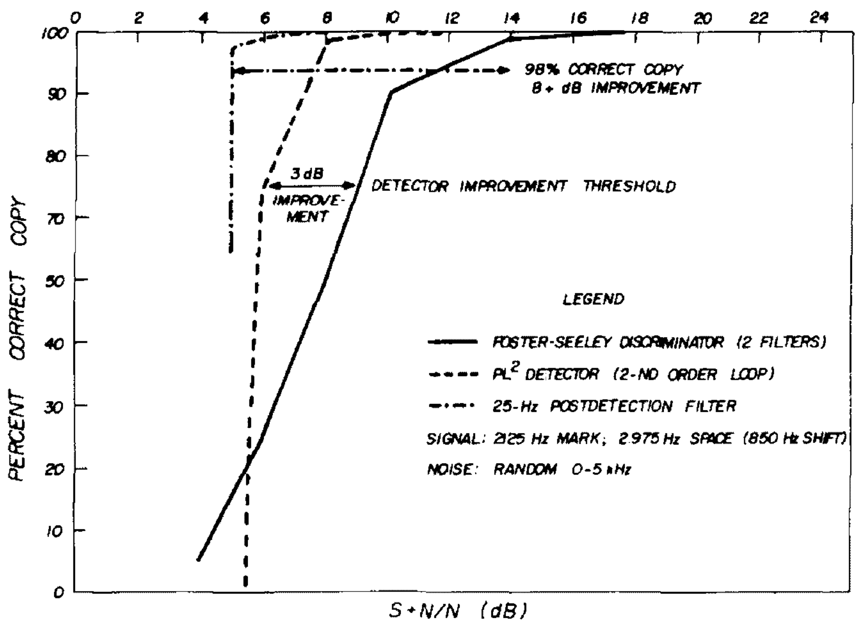


fig. 6. Comparison of the PL² and conventional Foster-Seeley discriminator.

design evaluation

The signal is now pure, has a square waveshape, and no noise exists except for a small amount that's within the same frequency spectrum as the desired signal. If this noise isn't of sufficient amplitude to trigger the pulse regenerator, the pulse won't appear in the regenerator output. The RC low-pass filter 6-dB point, between the tracking comparator and the Schmitt trigger, is at 25 Hz. If 100 wpm operation is desired, the capacitors in the filter (C15; C18 in **fig. 5**) should be .022 μF.

The PL² afsk detector has now met all design objectives:

1. Simple circuitry.

2. Copies all normal fsk shifts.

3. Mark-hold feature on random noise.

4. Works with low-frequency tones for ssb receivers and transceivers at the flip of a switch.

5. Afc circuit allows drifting or mistuned signals to be received with high-accuracy printout.

6. Tracking comparator provides anti-space function as the PL² shifts to the space frequency. If no shift, tracking comparator reverts to mark-hold mode.

TU performance data

Tests and measurements verified that the PL² provided 98% correct copy with

less than 6 dB s+n/n at the input, where noise was Gaussian from 0-5 kHz. **Figs.** 6 and **7** respectively show response curves and the test instrumentation used to obtain the data.

Approximately 8 dB improvement was obtained over a Foster-Seeley discriminator. The only disadvantage of the PL² detector was its wide bandwidth. The operational bandwidth is equal to the width, in Hz, of the fsk being received. The noise bandwidth is equal to the capture range, or about 1200 Hz. If an interfering signal 5-6 dB stronger than the desired signal falls in the PL² operational pass-band, it will capture the vco much like the fm capture effect. Other than this one minor detriment, everything is in the PL² afsk detector's favor.

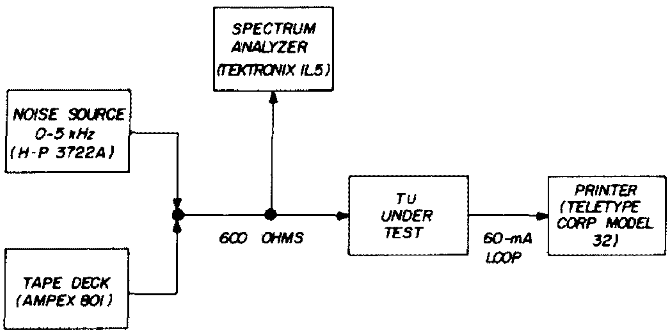


fig. 7. Instrumentation used to obtain data plotted in fig. 6. Afsk tones from tape deck, masked by noise, were received by the TU and printed on TT machine.

The vco center frequency can be set by an external resistor, so two pots and a switch were used. One pot sets the vco to 1500 Hz for use with ssb receivers and transceivers having limited audio pass-bands, thus the TU will operate on tones from 800-2200 Hz. The other pot sets the vco for 2550 Hz for normal RTTY tones, and the TU will operate on tones in the 1800-3200 Hz range. This made an afsk detector that is compatible with *any* communications receiver used for RTTY.

selector-magnet driver, autostart, and fsk

During the initial development of the PL² afsk detector, I used a Teletype Corporation Model 32 machine. This machine has a built-in solid-state SMD, loop-current sensing amplifier, and power supply. The SMD in the Model 32 was driven with a 12-volt, 60-mA loop supply directly from the Schmitt trigger output, with a 150-ohm resistor in the 12-volt supply to set loop current to 60 mA. (This resistor was also connected to the Schmitt trigger output via the TT loop.)

Most RTTY enthusiasts have TT machines that require an external dc-loop supply. These supplies have outputs ranging from 24 to 200 Vdc. The higher the voltage, the better will be the rise and fall times on the selector-magnet keying pulses. Thus the machine will operate on keying pulses with a wide range of distortion.

I wanted my selector-magnet driver transistor to key the loop at or very near ground potential. Also I wanted the stage to operate as a constant-current regulator so the keying-pulse waveform would be flat. I used a simple pnp transistor predriver (fig. 8) to develop forward bias across a pot for the SMD. This pot allows small changes to be made in loop current if TT machines are used that have different selector-magnet coil resistances.

Note that loop-voltage drop, when the loop is keyed or in mark, is obtained across an external resistor (R_{ext} in fig. 8). This protects the driver transistor from thermal overload.

autostart

The keyboard is connected in series with the SMD transistor emitter to ground. Isolation diodes connected to the junction of the SMD transistor emitter and keyboard allow keying of other circuits (fsk, afsk, etc.) with the keyboard. Such circuits are not disturbed when the loop is keyed in the receive mode by the SMD transistor, because the keyboard contacts are normally closed. This feature is important if you're using a transceiver with only a single vfo and you have it hooked up with the fsk circuit. When operating in this manner, it's necessary to offset receiver tuning with the incremental-tuning control to obtain the

afsk tones for the TU. I wanted a simple autostart circuit that would respond only to a fsk signal, and which wouldn't respond to a steady mark tone or a slow transition from mark to space and back at trolled triac, called a "Magtrac," whose reed-relay coil is IC compatible. This device will drive TT machines with motor-current requirements of 3 amps running and 12 amps starting, which

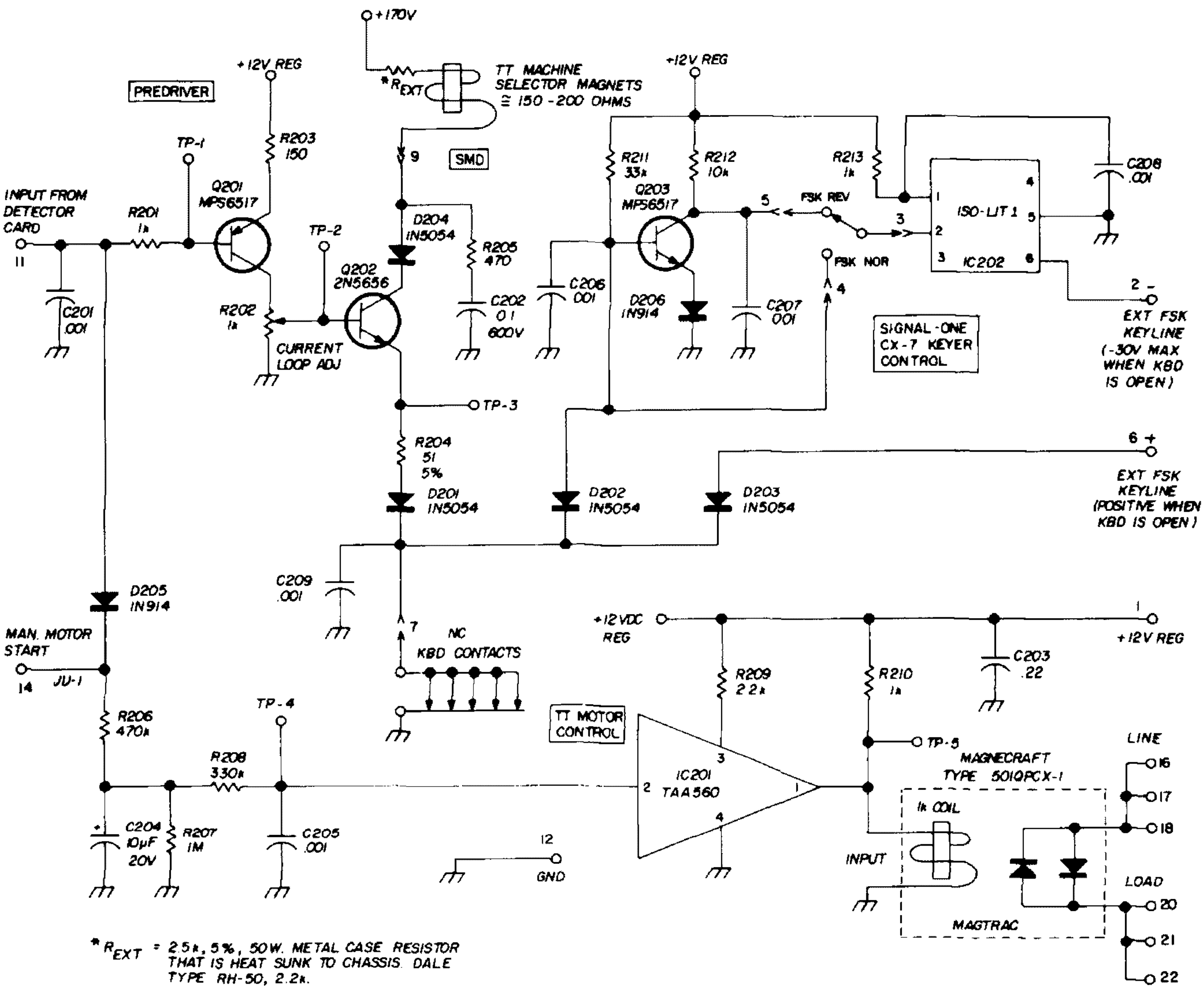


fig. 8. Schematic of selector-magnet driver and autostart circuits. Patent pending by WCI, 1971.

a slow rate, as is the case when someone is checking his fsk shift.

motor control

In my original circuit, the Schmitt trigger output was also fed through an isolation diode to a voltage divider and RC network. The output from the network would control a relay driver. However, I didn't want to use a large contactor relay in my final design. A trip to Magnecraft Corporation proved fruitful in this regard. They had a reed-relay-controlled includes any Teletype Corporation printer motor in amateur service. The schematic is shown in **fig. 8**.

In my final design (**fig. 8**) I used the TAA-560 to drive the Magtrac because the TAA-560's high input impedance doesn't load the RC circuit. In fact, I had to add a discharge resistor to obtain proper autostart turn-off time. The Magtrac offers great isolation between input and output while allowing a low-level logic signal to control a fairly high ac load.

Signal-One provisions

Out of necessity I added a circuit to key a negative voltage to operate the transmit fsk keying provision of the Signal-One CX-7 transceiver. (Several minor modifications are needed on the CX-7 for RTTY operation; but properly modified, the CX-7 works beautifully on RTTY.)

This circuit is also shown in **fig. 8**. A Litronix Iso-Lit 1, an electro-optical isolator, is used, which is a LED-control-led pnp photo transistor. A single transistor drives the Iso-Lit and also conditions the fsk signal so that mark is high in frequency on transmit and space is lower in frequency. The amount of shift is set by the front-panel fsk control on the CX-7.

construction notes

When building the PL² TU I didn't skimp on parts. Only the highest-grade commercial-quality components were used. All resistors are 5%, and capacitors are 10% tolerance. I recommend MIL-spec low-leakage electrolytic capacitors. Stick with the *exact* component values and *don't substitute!* Also don't use pull-out transistors, ICs, and diodes—use new devices that meet manufacturer's specs. You'll save yourself a lot of headaches.

The TU occupies two 4 x 4-inch plug-in PC boards. Standard 22-pin PC connectors are used, which fit a Cinch-Jones type 50-22A-20 socket. The entire main frame (**fig. 9**) of the RD-100 was built on a Bud AC-407 chassis. The unit is enclosed in an LMB CO-2 cabinet and front panel. The cabinet and panel, finished in two-tone gray, is a dead ringer for a piece of the S-line series.

The 12-volt dc power supply should be extremely well regulated with better than 0.1% regulation. I used a commercial unit with excellent performance, a Lambda Electronics Corp. Model LZS-11; quite a buy for only $38.00.

The loop supply is similar to that used by *W6FFC in his ST-6—nice and* healthy. Voltage is about 170 Vdc, and the supply provides better than 250 mA, of which only 60 mA are used. But the voltage stays put! Even if the voltage varies slightly, the constant-current characteristic of the SMD holds the loop at 60 mA on mark.

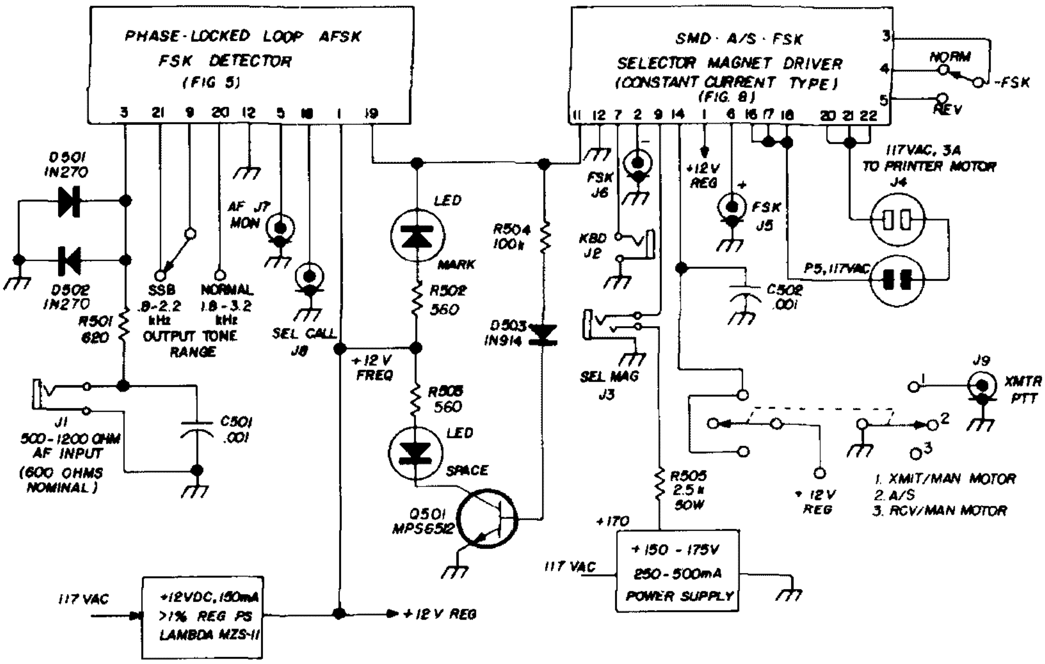
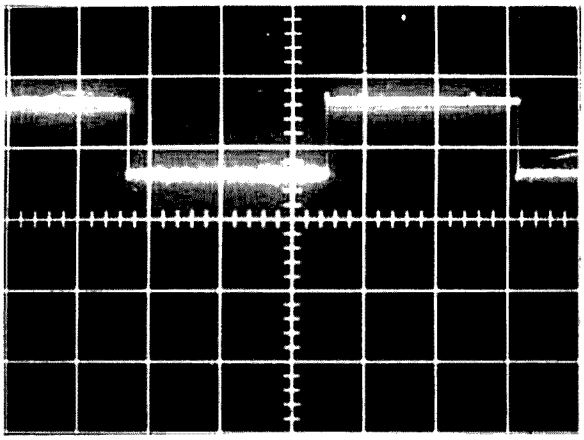


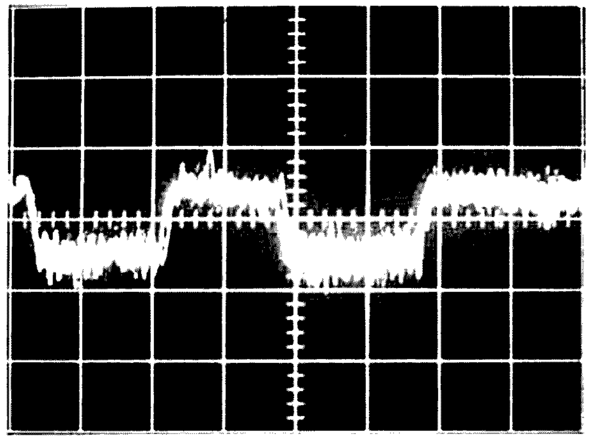
fig. 9. Main frame of the RCT-2D terminal unit. Patent pending by WCI, 1971.

fig. 10. Oscillograms showing response of PL^2 and SMD circuits at various test points.



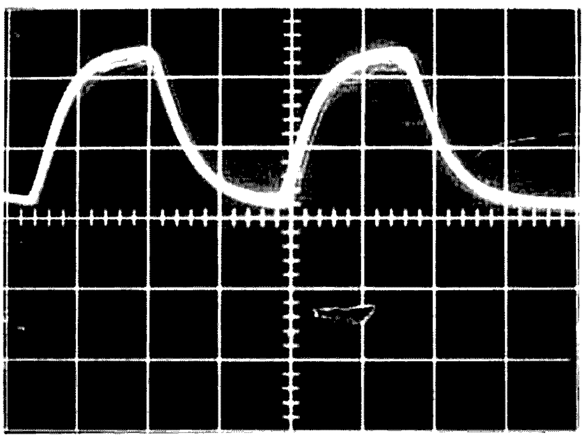
PL^2 vco output at TP2 on detector card (frequency test point).

horiz	10 ms/cm
vert	5 V/cm
F_{co}	1500 Hz
abscissa	0 Vdc



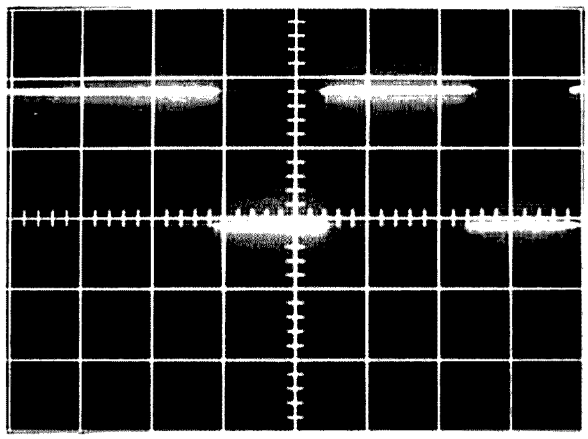
PL^2 demodulated afsk output at TP3 on detector card.

horiz	10 ms/cm
vert	100 mv/cm
ordinate	ref voltage at non-inverting input on IC2 (fig. 5)
shift	170 Hz
speed	60 wprn, 5-level



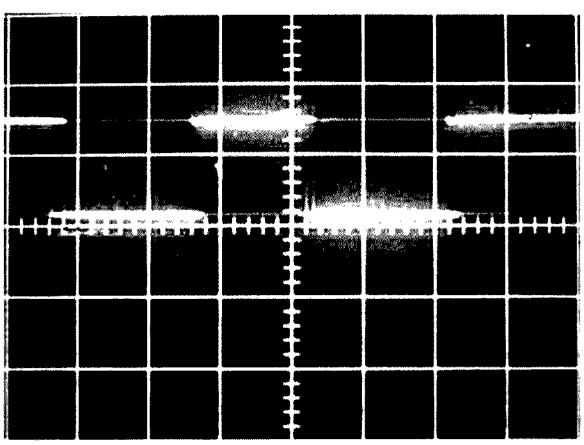
Tracking comparator output from low-pass filter at TP4 on detector card.

horiz	10 ms/cm
vert	1 V/cm
abscissa	0 Vdc



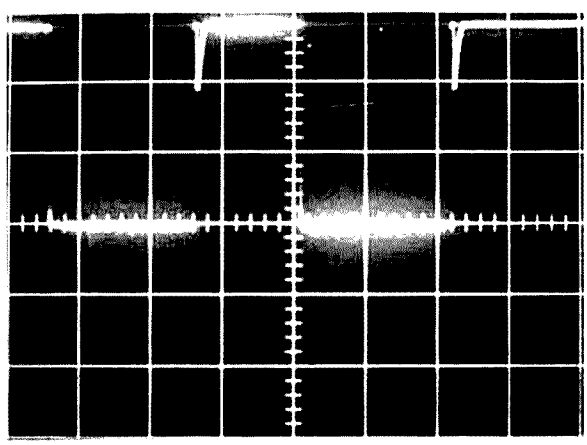
Schmitt trigger output at TP5 on detector card. (Mark low/space 10 Vdc.)

horiz	10 ms/cm
vert	5 Vdc/cm
abscissa	0 Vdc



SMD predriver output at TP2 on SMD card.

horiz	10 ms/cm
vert	5 V/cm
abscissa	0 Vdc



SMD output-current pulse (60 mA) to selector magnets.

horiz	10 ms/cm
vert	1 V/cm (or 20 mA/cm)

Resistor R_{ext} (**fig. 8**) is mounted directly on the chassis, which provides a heat sink. Be sure to use a small amount of Wakefield thermal compound to ensure good thermal contact to the chassis.

All input and machine-connection jacks are mounted on the chassis rear deck. The selector-magnet jack and plug are three-circuit types; however, only two circuits are used, as both sides of the selector-magnet coil are above ground. A grounding-type ac receptacle is also mounted on the chassis rear. This is the ac output line from the autostart, and the machine motor is plugged in here. Both power-supply primaries are fused. The 12-volt load of the unit is 50-60 mA.

initial setup and checkout

For the setup you'll need a vom and a frequency counter or oscilloscope and a calibrated af oscillator. These last two items can be used to read frequency in place of a counter. The oscilloscope can also be very helpful should you have to go "pulse hunting." The oscillograms of **fig. 10** will aid in alignment and checkout of both units.

1. Plug in the machine motor, keyboard, and selector magnets. Plug in the ac power to the RCT. Don't connect your receiver to the RCT at this time. Function switch should be in autostart; mode switch in ssb.

2. Connect a counter to TP-2 of the detector board and make sure that the mode switch is in ssb. Adjust R5 for a counter reading of 1500 Hz. This is the vco center frequency adjustment for ssb mode.

3. Place the mode switch in *NORM* and adjust R6 for a counter reading of 2500 Hz. This is the vco center frequency adjustment for the *NORM* mode.

4. Now connect your receiver 500-600 ohm audio to the RCT input. Set the volume at a comfortable level and tune to a clear frequency. The noise may trigger the autostart *ON*, but don't worry. Now adjust noise-squelch pot R16 to the point where the machine stops printing on impulse noise; ten seconds later the autostart will turn the motor *OFF*.

5. Place a 20-k/V vom or vtvm from TP-3 on the SMD board to ground, and adjust the loop-current-adjust pot for 3.75 Vdc. You now have 60 mA in the loop.

6. If you're using an ssb receiver or transceiver, enable the ssb mode and place the receiver/transceiver in lsb. Place the RCT in the ssb mode and tune to a RTTY station so that you can hear both tones. The RCT will do the rest.

If you're using a receiver with an af passband up to 3 kHz, then you can use the *NORM* mode on the RCT.

sources

1. The PL[2] (NE565K) and the tracking comparator (N5741T) are available from Signetics Corp., 811 East Arques Ave., Sunnyvale, Ca. 94086. The NE565K is approximately $11.00; the N5741T is approximately $1.50.
2. The Schmitt trigger, type TAA-560, is available from Amperex Electronics Corp., I. C. Division, Providence Pike, Slatesville, R. I. 02876. The TAA-560 is approximately $3.00.
3. The Magtrac type 501QPCX-1 is available from Magnecraft Electric Co., 5575 N. Lynch, Chicago, Ill. 60630 for $16.80.
4. The 12 Vdc regulated power supply, model LZS-11, is available from Lambda Electronics Corp., 515 Broad Hollow Road, Melville, L. I., New York, 11746 for $38.00.
5. The Iso-Lit 1 is available from Litronix Corp., 19000 Homestead Road, Cupertino, Ca. 95014, for about $4.00.

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3. Application Note D41, LIN-023-110, "10M Phase-Locked Loop 565," Signetics Corp.
4. Application Note D45, LIN-010-80, "35M High-Performance Operational Amplifier 5741," Signetics Corp.
5. Application Note "Type TAA-560 Level Detector/Schmitt Trigger," Amperex Electronics Corp.
6. Application Note "Iso-Lit 1, Electro-Optical Isolator," Litronix, Inc.

ham radio

introduction to microwaves

A discussion of the differences between low frequencies and the microwave domain in terms of ac-circuit theory

William L. Sullivan, W1CBY

The amateur bands above 1 GHz offer interesting challenges to the builder-experimenter and operator alike. Known as the microwave region, this portion of the radio spectrum is relatively unpopulated with amateur signals when compared to the lower-frequency bands. Many amateurs don't understand the theoretical and physical concepts of microwave work and tend to ignore this interesting facet of ham radio. Many of the devices used at the microwave frequencies were developed in laboratories by physicists; consequently the behavior of such devices is usually described in complex mathematical terms. However, by using a logical down-to-earth approach, microwave fundamentals are easily understood.

50 MHz and up

Construction technique and operating know-how in the region above 50 MHz is much different than in the "dc bands." Getting equipment to work at, say, 1296 MHz is a real test of one's ability in circuit design techniques and testing. If you like to build equipment, here's a fertile field for your ingenuity and patience. Like to work DX without the hassles and pile-ups of 20 meters? Try over-the-horizon work with narrowband equipment on 3300 and 5650 MHz.[1] Try

some of the experiments with ROCLOC (relative or crystal local oscillator control), a technique developed by the San Bernardino Microwave Society.[2]

Many other activities will be found in the amateur bands above 50 MHz. Communications by tropo scatter, meteor scatter, and moonbounce are examples. Recent work by amateurs has established the practicability of communications over long paths on frequencies to at least 10,000 MHz. Interested in building equipment? Here's a remark by an East-coast microwave enthusiast, K2JNG: "You don't buy gear for this band (1296 MHz), you build it."[3]

These are but a few of the happenings in the bands above 50 MHz. Now let's look at some of the differences in techniques and technical disciplines required for microwave work as compared to those associated with the lower frequencies.

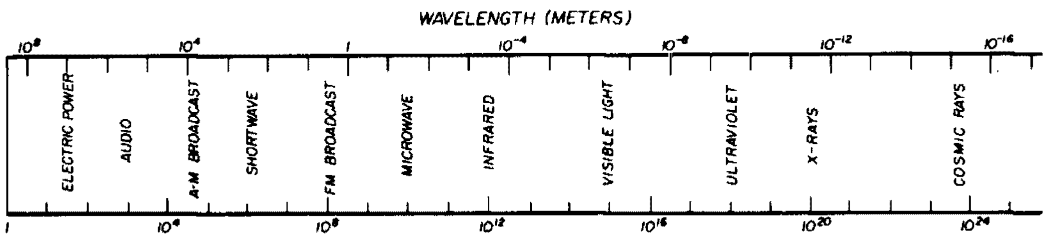

fig. 1. The electromagnetic spectrum.

background

Electricity that alternates with time, whether from power mains, radio-broadcast stations, or radar is described in terms of an alternating wave. It makes no difference whether the circuit involves vacuum tubes, transistors, transmission lines or antennas; the resulting phenomenon may be attributed to a series of waves. Scientists recognized the similarity between these various waves before the turn of the century and segregated them by frequency as shown in **fig. 1**. For reasons we will see shortly, this band of frequencies is called the electromagnetic spectrum. All waves within the electromagnetic spectrum travel at the speed of light — 300 million meters/sec or about 186,000 miles/sec. If the operating frequency is 10 billion cycles/sec (10,000 MHz) the alternating wave will go through one complete cycle in one ten-billionth of a second. Since the wave moves at 300 million meters/sec, it will move one ten-billionth of this distance during one cycle, or 3 cm. This distance is defined as the wavelength (**fig. 2**). The point is, although the speed of light may seem instantaneous, an electromagnetic wave requires a finite amount of time to move from one point in space to another. When the frequency is low, it takes longer to complete one cycle, and the wavelength is longer. At frequencies above 1000 MHz, however, the wavelength becomes relatively small, and this part of the spectrum is called the microwave region.

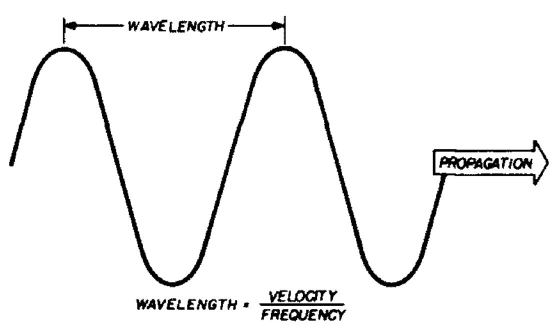

fig. 2. The wavelength equation.

The increasing use of microwaves in recent years parallels the expansion to the hf and vhf portions of the spectrum of a few years ago. In all cases the spectrum has become crowded with radio services, and it has been necessary to expand in the only possible direction, upward. Measured in cycles, the additional fre-

quencies made available by the advent of microwave technology are several hundred times greater than all those previously available. These additional frequencies are not a panacea, however, because many of the new services require wide bandswidths. Radar, for example, requires a handwidth of 5 to 10 MHz.

bandwidth considerations

The microwave frequencies are especially important for two reasons, both based on the peculiarities associated with ultra-short wavelengths. The first is the bandwidth consideration; a radio signal consisting of only one frequency isn't practical for the transmission of intelligence, because a single-frequency signal is infinitely narrow. Information may be carried by a signal only if it occupies a restricted band of frequencies. In general, the wider the band, the more information can be conveyed per unit time. The intelligence may consist of telegraphic dots and dashes, audio or video modulation, radar pulses, etc. In each case a certain bandwidth must be assigned to provide distortion-free transmission and reception.

When the operating frequency is very high to start with, a given bandwidth constitutes a smaller percentage of the carrier frequency, and it's possible to provide more channels of intelligence. For example, if we modulated a 10-MHz radio signal with video 2 MHz wide, the bandwidth would constitute 20% of the carrier frequency. If, on the other hand, we modulated a 100-MHz signal with the same video information, the resultant bandwidth would be only 2%. In general, it's difficult to obtain bandwidths greater than about 5% of the carrier frequency.

Bandwidth becomes increasingly important when the modulating signal is characterized by an inherently wide bandwidth. This is the case with the very short pulses associated with radar and pulse-code communications systems, or where the multiplexing of a large number of video channels by a single transmitter results in economic television relay service.

antennas

The second major consideration has to do with antennas. Both the beamwidth and the gain of an antenna depend upon the ratio of the wavelength to the size of the antenna. In many applications, it's desirable to concentrate the radiated energy into very narrow cones a few degrees wide, with gains of several thousand. This can be done only when the antenna is extremely large in terms of operating wavelength and is impossible when the wavelength is measured in miles. However, extremely high-gain antennas are practical with wavelengths less than an inch.

atmospheric noise

Very low atmospheric noise exists in the microwave region. A "low-noise window" extends from 200-1500 MHz where atmospheric noise is almost non-

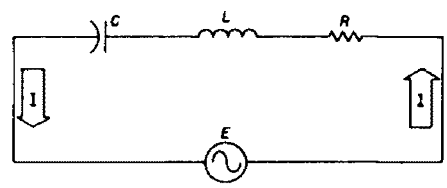

fig. 3. Concept of displacement current in an ac circuit.

existent. Additionally, the very short wavelengths associated with microwave frequencies are suited to many unique construction techniques. Hollow pipes may be used for transmission lines, cavity resonators with extremely high Qs may be used in place of LC circuits, and parabolic reflectors may replace massive directive antenna arrays.

electrical concepts

The misunderstanding and confusion surrounding microwaves is due in part to the apparently different electrical concepts that govern their operation. Actually the concepts are not so very different from those encountered at lower frequencies; it's just that the laws used at the lower frequencies are not sufficiently

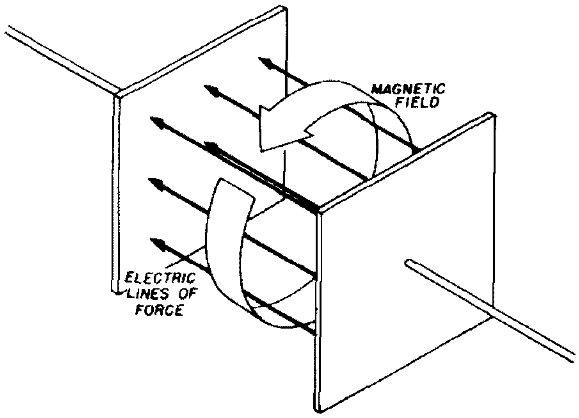

fig. 4. Electromagnetic field between capacitor plates.

general to encompass microwave circuits. This difference is comparable to the difference between dc and ac circuit theory. After studying dc circuits, it's necessary to have an understanding of frequency, phase, and reactance before ac circuits may be properly analyzed.

In dc and low-frequency communications, it's customary to consider the transmission of energy through the flow of current or electrons. In the microwave region this concept may be misleading, because it doesn't explain the transmission of energy through a hollow waveguide where conventional current does not exist. If, instead, we consider that energy is contained in an electromagnetic field, the transmission phenomenon at *any* radio frequency may be easily explained.

Many tend to forget the basic elements of static electricity and charges once the concepts of voltage, current, and Ohm's law are understood. This isn't too unusual, because these latter units are easily measured in everyday electronics circuitry. However, at microwave frequencies, the familiar quantities of voltage and current are difficult, if not impossible, to measure, and we must resort to the fundamental concepts of electric and magnetic fields to understand microwave theory.

electrical fields

Before progressing into a general discussion of fields, let's consider the concepts of electric fields and electric voltage. Actually, they are very closely related quantities, with field strength expressing the force that would be felt by an electron placed at some point in space and voltage expressing the product of this field strength and the distance over which it moves the electron. Contrary to what you might think, this is not a new idea at all. If you'll recall your first studies in electricity, you'll no doubt remember the experiments with cat fur and a rubber rod. When you rubbed the fur with the rod, you generated a static electric field. The electric field associated with radio waves differs from the electrostatic field *only* in that it varies in time; this is not unlike the difference between ac and dc.

We know, from the principles of continuity, that the current leaving the signal source E at the right in **fig. 3** is exactly equal to the current returning at the left. The current flows through the resistor and the inductor, but what happens at the capacitor? There is no movement of electrons through the dielectric separating the plates of the capacitor, so there appears to be a discontinuity in the current flow. However, from our experience in low-frequency ac circuits, we know that current apparently flows through the capacitor.

displacement current

This difficulty was recognized many years ago and was straightened out by inventing a new kind of current called displacement current. Actually, modern physicists do not feel that displacement current is so much a current as did the early researchers; nevertheless the concept is still valid. In fact, since displacement current produces the same effects as actual conduction current in describing this phenomenon.

The displacement current between capacitor plates is a function of the changing electromagnetic field between the plates (**fig. 4**). In elementary static-electricity theory, displacement current doesn't exist because the electrostatic field doesn't vary with time.

the flux equation

The magnetic field surrounding a current-carrying conductor may be expressed by

$$H = \frac{I}{2} r \qquad (1)$$

where H is the magnetic flux density at a distance r from a wire carrying a current I (fig. 5).

Michael Faraday (1791-1867) discovered that an electromotive force E could be generated by a magnetic field cutting the current-carrying conductor:

$$E = \frac{H}{t} \qquad (2)$$

where the induced voltage E is proportional to the magnitude of change in the magnetic flux and inversely proportional to the time in which the flux changes.*

From (1) it can be seen that the induced voltage E will increase with either a larger or faster change in magnetic flux density H. In other words, if the flux change is held constant, the induced voltage will increase with the frequency of the magnetic-flux change.

the electromagnetic field

Note that no mention is made of either conductor size or material in either of the above equations. If the size of the conductor were gradually reduced step by step, the induced voltage would remain the same, because the law states nothing to the contrary. In fact, the size of the wire may be reduced until there is no wire, but there will still be an induced voltage across the space that was occupied by the wire.

The voltage induced across space by the changing magnetic flux is in the form of an electric field. This field causes the flow of displacement current; the displacement current generates a magnetic

*The classic definition of the relationship between magnetic flux and magnetic flux density is complex. An introduction to these and other elements of magnetostatics appears in reference 4. *Editor.*

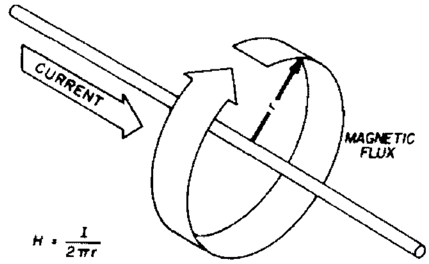

fig. 5. The flux equation, which describes the magnetic field surrounding a current-carrying conductor.

field in direct proportion to its intensity. The electric and magnetic fields then are complementary, with one generating the other as they move through space at the speed of light. The combined effect of electric and magnetic flux form an electromagnetic field.

The electromagnetic field may be represented as in **fig. 6**, where the electric and magnetic lines of force are perpendicular to each other and mutually perpendicular to the direction of propagation. Although the electric and magnetic fields are represented by lines of force, separate E and H lines don't exist in space; rather a kind of electric or magnetic tension exists that's directly proportional to the intensity of the flux. That is, more lines of force are used to illustrate a strong field than a weak one.

the microwave domain

The fact that a changing magnetic field induces a voltage across space is the concept needed to understand microwave circuits. In other words, at ultra-high frequencies the current flows not only in conductors; it may flow as displacement current between points in space. In a sense there is still a circuit, but the paths are no longer well defined.

One of the outstanding events in the history of microwave theory was the recognition of the similarities between the effects of different electromagnetic field configurations and those in inductive, capacitive, and resistive elements – the circuit components familiar at the lower frequencies.

When the basic circuit elements are viewed from the fundamental concept of energy, it becomes obvious that the

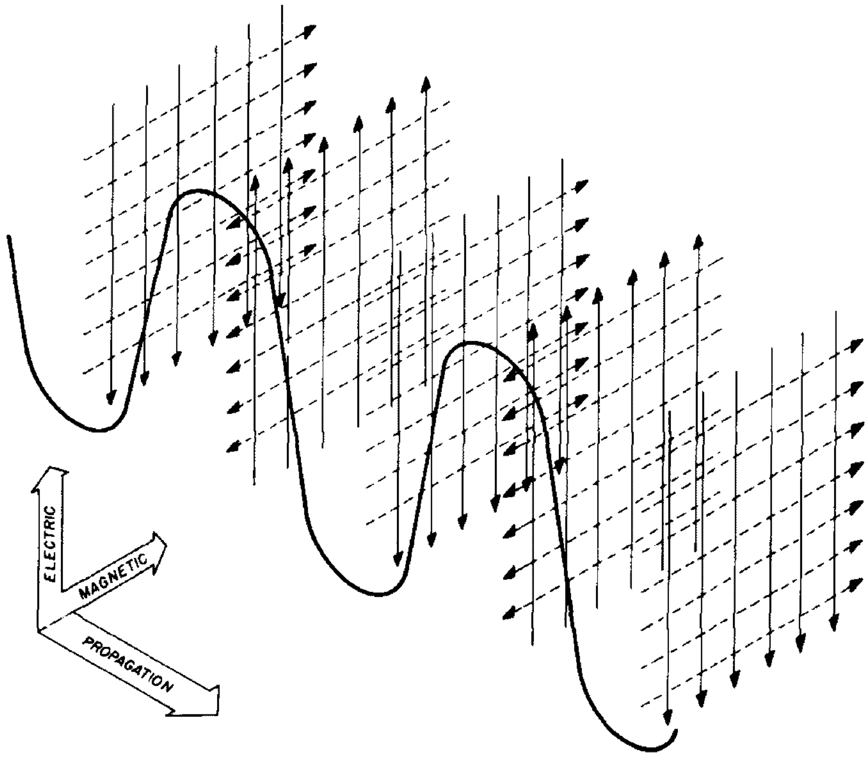

fig. 6. Electromagnetic field associated with an electrical wave varying in time, magnitude, and direction.

energy of an inductor is always stored in a magnetic field, and the energy of a capacitor is always stored in an electric field. Therefore, any microwave component that stores energy by the virtue of an electric field is regarded as having capacitance. Conversely, any component storing energy in a magnetic field is considered to possess an equivalent inductance. When microwave energy is dissipated in terms of heat, the effect is similar to a resistive component in a low-frequency system.

With certain reservations, the impedance associated with a microwave circuit may be related to the R, L, and C elements in exactly the same way as at the lower frequencies. In a low-frequency circuit, where the wavelength is many times larger than that of the individual circuit elements, no anomaly exists. However, in a microwave system, the wavelength is comparable to *each* of the circuit elements, and the magnitude of the R, L, and C components is dependent upon wavelength. Therefore, an inductive or capacitive value may be assigned to a microwave component only when the *frequency is held constant.* However, in practical circuits where bandwidth is relatively narrow, this concept allows one to use his knowledge of low-frequency ac theory to increase his understanding of microwave circuit with a minimum of error.

references

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2. Dick Kolbly, K6HIJ, "Standards for Amateur Microwave Communications," *ham radio,* September, 1969, p. 54.
3. Bill Smith, KØCER, "The World Above 50 Mc," *QST,* July, 1971.
4. Keith Henney, "Radio Engineering Handbook," McGraw-Hill, 1959, pp. 1-76 to 1-78. Also pp. 1-88 to 1-93 et seq.

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Guba and Zimmer, "Pulse: A Practical Technique for Amateur Microwave Work," *QST,* February-May, 1963. Also *VHF Manual,* Ch. 10.

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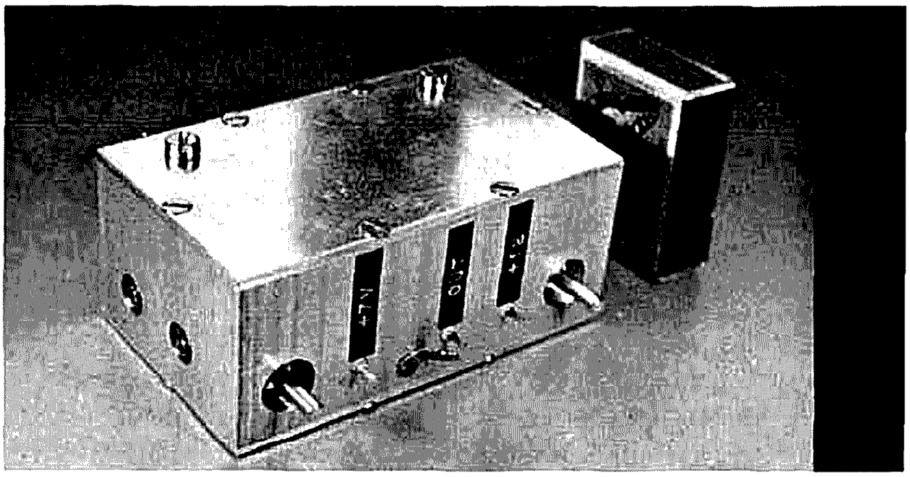


high-stability variable frequency oscillator

This high-frequency 5.0- to 5.5-MHz vfo provides excellent stability and maximum low-harmonic output with a minimum number of components

Veikko Aumala, OH2CD, Helsinki, Finland

The highly stable vfo described in this article has evolved over several years. It features sufficient output for any mixer configuration, is very solid, has small size and exhibits excellent spectral purity. It tunes the frequency range from 5.0 to 5.5 MHz with ±150 kHz overlap.

The basic oscillator circuit is the inherently stable Seiler type. In the Seiler circuit the transistor and tank circuit are very lightly loaded, resulting in excellent frequency stability. The tuning capacitor is a small type used in fm broadcast receivers; it has a built-in 3:1 spring-loaded drive mechanism. With the dual tuning capacitors used in this design (see fig. 1) either one of two output frequencies may be selected by relay K1.

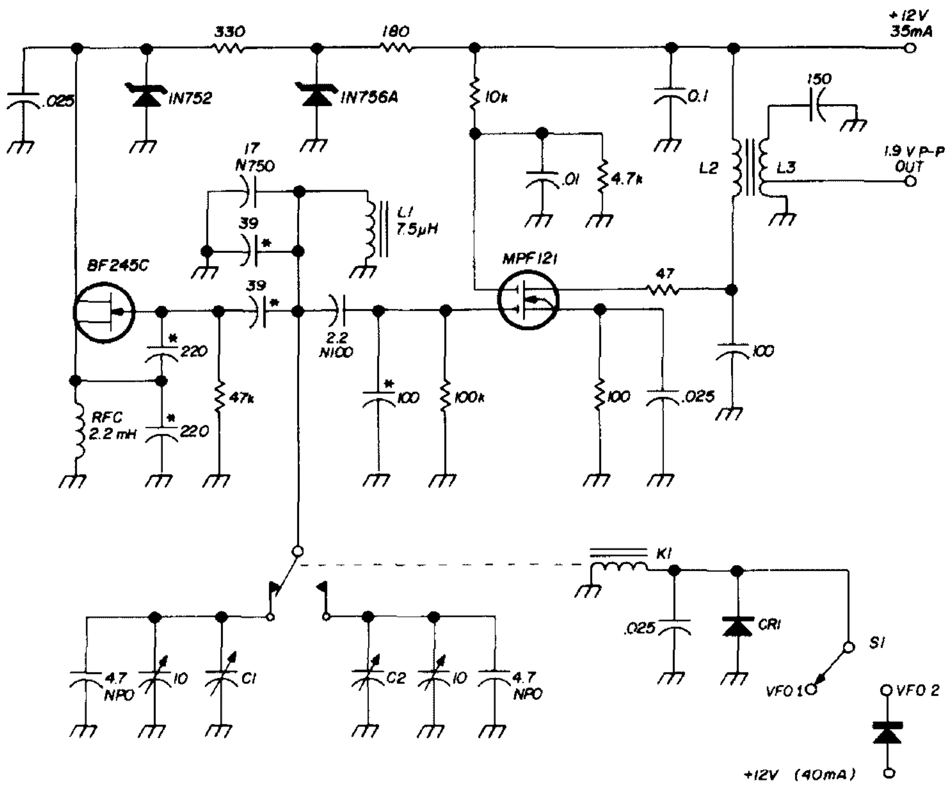
The vfo circuit is housed in a very solidly built aluminum box approximately 4 inches long, 2½ inches wide and 3/16-inch thick. This construction provides a high thermal integration constant so ambient temperature variations have little effect upon the oscillator frequency.

circuit

In the circuit in fig. 1 the lower limit of the variable tuning capacitance is approximated by

$$C = \frac{\Delta C}{\left(\frac{\Delta f}{f - \Delta f}\right)^2 - 1}$$

where C is the required minimum tuning



Capacitors marked with asterisk are mica; NPO capacitors are N750 ceramic types.

C1, C2	3-section 3.2 to 18.5 pF variable, parallel connected
L1	7.5 µH. 13.5 turns 0.5 mm enameled (no. 24) in 2 layers on ferrite pot core
L2	30 turns 0.2 mm enameled (no. 34), honeycomb wound, 4 mm (0.16 inch) long, on 65 mm (1/4 inch) slug-tuned form
L3	30 turns 0.2 enameled (no. 34), honeycomb wound, 4 mm (0.16 inch) long, on same form as L2 (see text)

fig. 1. Schematic diagram of the high-stability vfo. One of two output frequencies is selected by K1. If K1 is controlled by a multivibrator two receiving channels may be monitored at the same time.

capacitance, f is the upper frequency limit, Δf is the frequency change and ΔC is the total tuning capacitance variation.

The unused tuning capacitor has some affect upon the output frequency although it is isolated by the small capacitance (about 4 pF) of the open relay contacts. The greatest effect occurs at the upper frequency limit when the main tuning capacitor is at minimum capacitance. This effect can be minimized by making the minimum tuning capacitance as large as possible by sharing the fixed capacitors in the tuned circuit.

If you already have a stable regulated 12-volt supply you can eliminate the 1N756A zener diode and the 180-ohm resistor. Power consumption is 35 mA with the zener circuit, 25 mA without it.

Use a fast, reliable relay to switch the tuning capacitors. The rf voltage at this point in the circuit is 30 to 40 V p-p. A high-quality reed relay is recommended. The diode across the relay coil eliminates the negative voltage transient when the coil is de-energized.

If you want two-channel control for your receiver use a multivibrator circuit to control the relay. Try a multivibrator frequency between 10 and 300 Hz. In addition to the two desired channels there will be adjacent sidebands with spacings equal to the multivibrator frequency.

In a circuit of this type the vfo signal is usually taken from the source of the oscillator fet. On a good oscilloscope the waveform at this point is very close to a

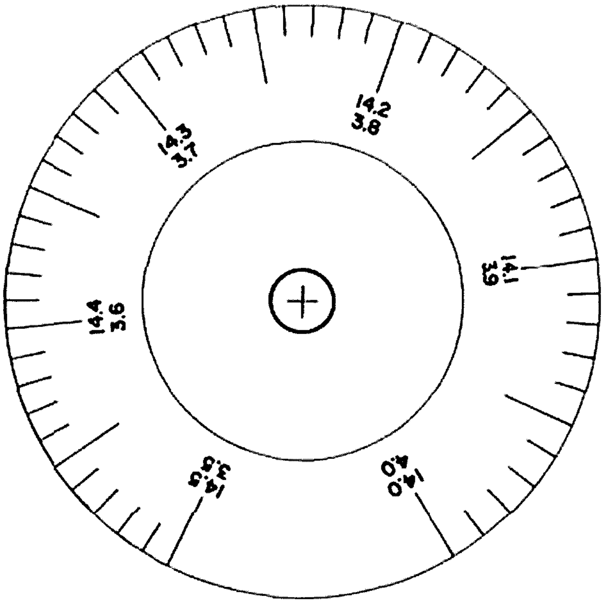


fig. 2. Full-size dial for the 5.0 to 5.5-MHz vfo.

true sinusoid. However, a check with a spectrum analyzer will reveal that the second harmonic is only 20 dB down. Since the output signal should be as clear of harmonics as possible, in fig. 1 the oscillator output signal is taken from the frequency-determining circuit. At this point in the circuit the second harmonic is more than 50 dB down.

The vfo input signal to gate 1 of the MPF121 amplifier should not be greater than 0.9 V p-p. The MPF121 is an inexpensive, zener-protected mosfet and will provide all the needed amplification with good isolation between input and output. The drain current of the MPF121 is set to 10 mA by selecting the source resistor; this current level results in maximum linearity.

The 47-ohm resistor in the drain circuit of the MPF121 prevents vhf parasitic oscillations. I found that the gate-1 lead inductance and drain lead inductance resonated with the 100 pF capacitors to form a tuned-drain, tuned-gate oscillator around 800 MHz. This parasitic oscillation is effectively damped by the 47-ohm resistor in series with the output tuned circuit.

Inductors L2 and L3 are both honeycomb wound. One of the coils may be moved along the coil form to control coupling. The output bandpass of this circuit should be tuned with a sweep generator to obtain optimum performance. Output from the sweep generator is terminated with a 50-ohm resistor and fed to gate 1 of the MPF121. The vfo output is connected to 50-ohm coax to the input of the sweep generator, terminated by a 75-ohm *resistor*. This is approximately a 30-ohm load. With this arrangement output is flat within 0.1 dB from 5.0 to 5.5 MHz; attenuation at 10 MHz is about 50 dB.

A hot-carrier diode ring mixer requires about 10 mW of rf injection, much more than other mixer arrangements. The output of this circuit is just sufficient.

When aligning the vfo be careful not to overdrive the MPF121; 0.9 V p-p is enough. More drive will not increase the desired output signal but will produce considerably more harmonics.

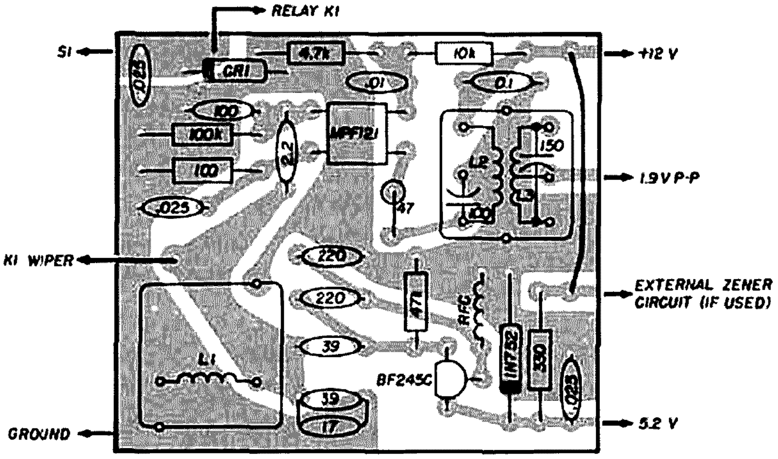
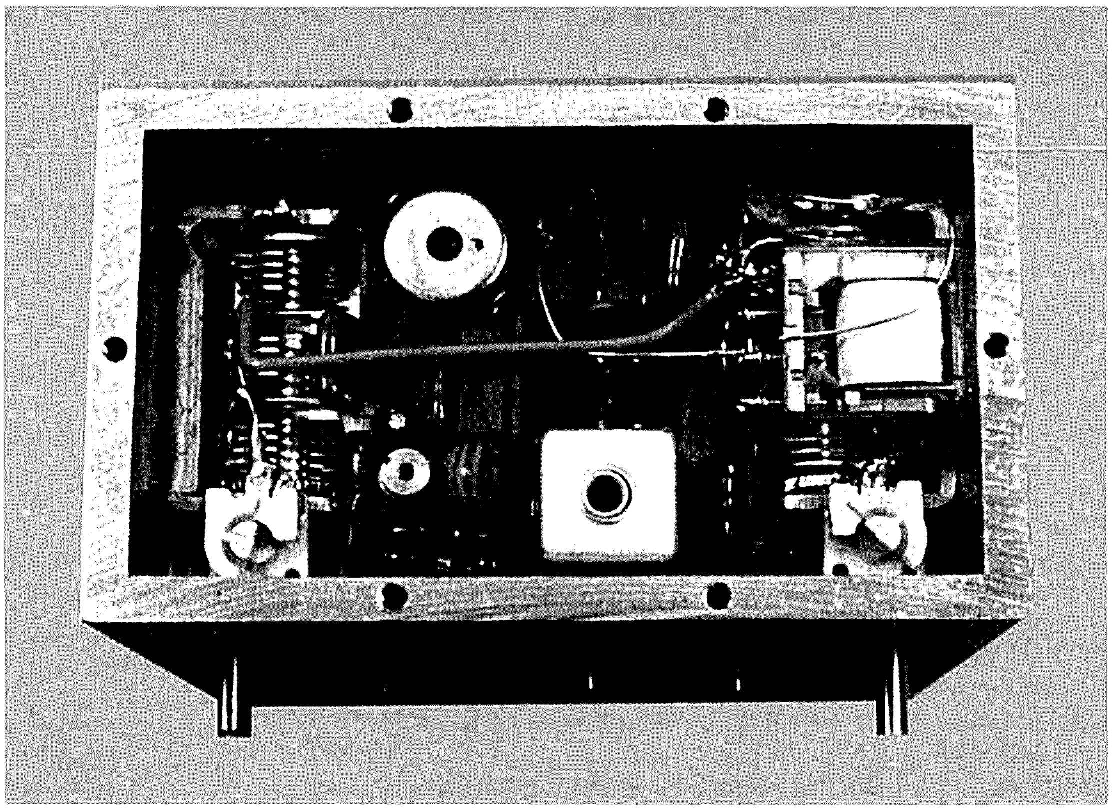


fig. 3. Full-size printed-circuit board layout for the vfo.



The high-frequency vfo is built into a rugged aluminum chassis. Layout is shown in fig. 4.

construction

The high electrical stability of this circuit cannot be realized unless the actual physical construction is mechanically stable. The temperature delay through the 3/16-inch sides of the cast-aluminum chassis shown in the photos is about ten minutes.

For maximum rigidity the two tuning capacitors are screwed to the sides of the chassis. Distance between shafts is 75 mm (2-7/8 inches); therefore, two 70-mm (2-3/4-inch) dials can be used (see fig. 2).

The printed-circuit board shown in fig. 3 is soldered directly to feedthrough inserts and grounded in two places with metal straps. Both sides of the board are easily accessible.

If a ferrite core is used in L1, the relay should be placed as far away as possible. The magnetic field around the relay coil affects the ferrite; magnetic shielding may be necessary.

The gear drives to the tuning capacitors are left outside the aluminum chassis. I have not found a single reliable planetary drive mechanism; spring-loaded gears are preferred. A ratio of 10:1 is sufficient; this provides about 50 kHz per turn. I found a 36:1 drive from Semcoset in Germany which gives about 15 kHz per turn.

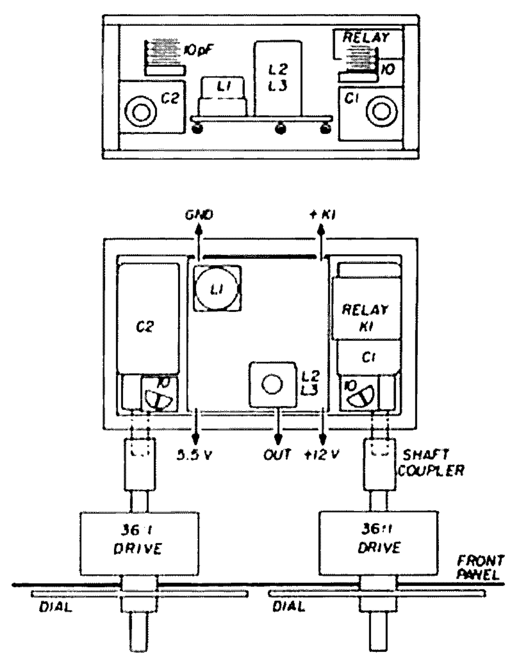


fig. 4. Mechanical layout of vfo.

table 1. Operating characteristics of the high-stability vfo.

frequency range	4.850 to 5.650 MHz
frequency used	5.000 to 5.500 MHz
output level	1.9 V p-p ±0.5 dB into 35 ohms
stability	15 Hz per °C 5 Hz per volt (9 to 20 volts)
output open/short circuited	20 Hz difference
harmonic content	more than 70 dB down
mechanical shock	less than 100-Hz frequency change with several successive hard shocks
supply voltage	10 to 18 V, 12 V nominal
power consumption	35 mA (relay, 40 mA)
size	100 x 60 x 46 mm (3.9 x 2.4 x 1.5 inches)

summary

The complete operating specifications of this vfo are given in table 1. Several similar units have been built, and all exhibit the same operating characteristics.

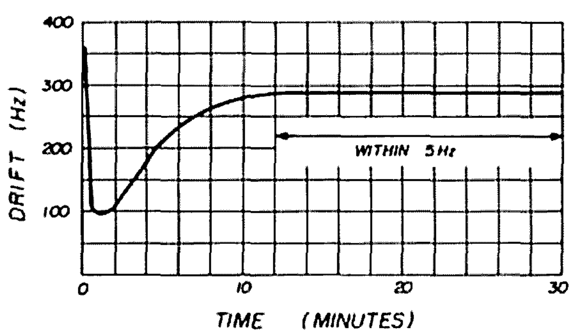
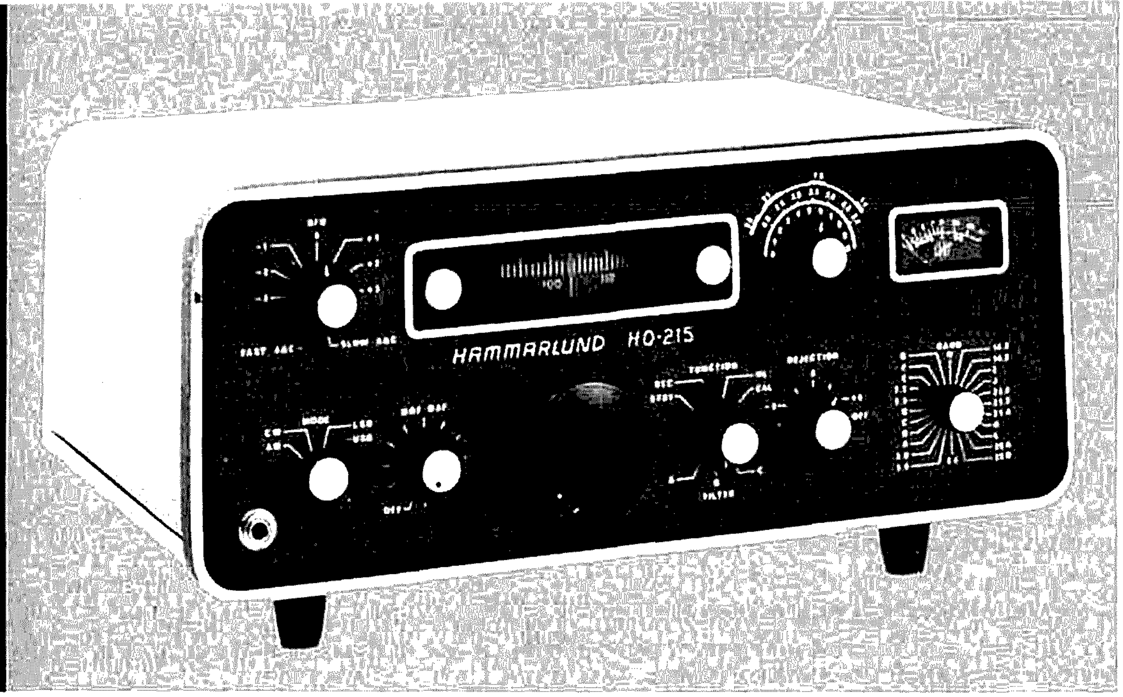


fig. 5. Frequency drift characteristics of the vfo when turned on from a cold start.

This is by no means an exceptional device, but it is a good answer for the modern amateur. When using a hot-carrier diode ring mixer in a bilateral transceiver this vfo uses the minimum number of componenets for the features it provides.

ham radio



adding 160-meter coverage to the HQ125 receiver

This simple modification should add rather than detract from the resale value of your HQ125. No metalwork surgery is required

Stuart Meyer, W2GHK, 4000 Wingate Drive, Raleigh, North Carolina

Although the Hammerlund HQ215 has a number of positions on the bandswitch for additional 200-kHz segment coverage on the various bands, the receiver's design does not permit such frequency ranges below approximately 3.4 MHz. After studying the problem of 160-meter coverage, the following relatively simple modification was devised. It can be accomplished with reasonable skill, without tearing apart any of the existing circuitry and without the need for drilling any of the metalwork.

An additional crystal is required to cover the range of 1.8 to 2.0 MHz. Crystals for several receivers modified by me were obtained from the International Crystal Manufacturing Company.* The appropriate pages of the Hammerlund manual were forwarded to International Crystal so the crystals would correlate and calibrate properly. The frequency of

*International Crystal Manufacturing Company, Inc., 10 North Lee, Oklahoma City, Oklahoma 73102.

the crystal required for this modification is 4.9550 MHz. The various fixed and trimmer capacitors called for at the end of the article should also be on hand when the modification is started.

Complete coverage of the 160-meter band is not possible with a single position of the band selector due to the relatively small value of available capacitance change compared to the large values to be added by this modification. Therefore, it is necessary to split the band segments into two segments of 100 kHz, each covered by separate positions (B and C) of the bandswitch, to obtain optimum sensitivity over the entire band. Tuning of the second oscillator stage, however, will cover the entire 160-meter band regardless of the setting of the bandswitch with respect to position B or C.

Two sections of instructions follow which will allow you to cover either segment as a complete operation, or, by completing both section, to cover the entire band.

It is assumed that the receiver to be modified is in normal working order and does not require service. Remove the top, bottom and the side panel nearest the bandswitch in accordance with the instructions in the HQ215 service manual. Note: The following procedures assume that the receiver is upside down on the workbench with the front panel closest to the operator.

1.8 to 1.9 MHz

Install a 4.9550-MHz crystal into position C (not B) of the crystal deck. Positions 5 and 6 of section D of switch S-101 should be shorted by soldering a short length of number-18 or 20 solid tinned wire across the switch tabs. Do not attempt to insert the wire into the eyelet holes, simply lay the wire on top of the tabs. Be sure that no solder droppings or bits of wire fall down into the receiver. Pins 5 and 6 of section D are on the wafer closest to the rear of the receiver, and in several units I have checked these pins have green and yellow wires.

Trimmer capacitor C127 (refer to figure 4-3 of the HQ215 manual) should be adjusted to its maximum capacity (maximum clockwise tension on the screw) but be sure to not apply so much tension as to strip the threads. When operating properly, it should be possible to hear a loud spurious signal between 3 and 5 kHz below 1800 kHz. This is a normal "birdie" for the frequency conversion scheme used in the HQ215. It may be necessary to parallel C127 with a 220-pF silver-mica capacitor if the resultant note at 1.8 MHz using the *calibrate* position is not clean. Readjust C127 to obtain the cleanest sounding marker signal.

Although the schematic in the manual shows trimmer capacitors C120, C125 and C130 connected to contact 6 of S101A, B and C, respectively, a physical examination of the bandswitch reveals that they are actually connected to terminal 5, with jumpers to 4 and 6. Very carefully cut the jumpers between terminals 5 and 6 of the three front (A, B and C) sections, leaving terminal 6 clean and ready for the additional padders to be installed. Extreme care must be used at this stage of the conversion to insure that the bandswitch is not damaged and that no solder droppings fall into the receiver. If this does happen, they should be shaken loose and removed completely before proceeding.

Next, prepare the three variable trimmers (310-700-pF, Electromotive type PC4215) as follows: For switch section A, wire a 1000-pF silver-mica capacitor across the variable trimmer. For switch section B, connect a 1500-pF across the trimmer, and for section C (the first mixer circuit) use a 1000-pF capacitor across the trimmer. The leads on the fixed capacitors should be carefully trimmed and wrapped around the tabs of the trimmers and soldered to hold the leads near the base of the tabs.

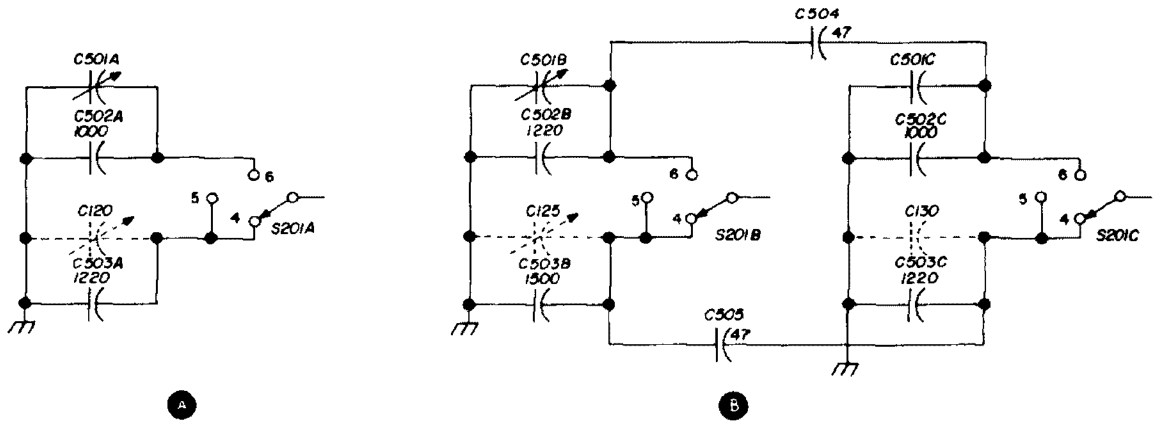
Before installing the new padders, place them in approximate position and note the spot on the grounding shield between switch sections where one side of the trimmer will anchor. Tin this area

of the shield and then install the padders between the appropriate pins and the tinned spot on the shield. Be careful not to force the switch tabs out of their normal position so there will be no strain on the switch when this operation is completed.

One last operation is left. Install a 47-pF silver-mica capacitor from terminal 6 of S101B to terminal 6 of S101C. This MHz, with the exception of alignment.

1.9 to 2.0 MHz

Prepare two 1220-pF capacitors by wiring a 1000-pF and a 220-pF silver-mica in parallel. Install one of these 1220-pF capacitors across C120 (switch section A) in *Section 1.* Install a 1500-pF silver-mica capacitor across C125 (switch section B) and the other 1220-pF unit



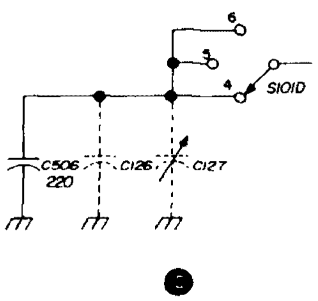


fig. 1. Simple circuit modifications put Hammarlund HQ215 receiver on 160 meters. Higher values of fixed capacitance are used on the higher-frequency 160-meter segment (1900 to 2000 kHz) because of the lower value of trimmer capacitance which is factory installed at position 5 of the bandswitch.

C501A,C501B, C501C	310 to 700 pF trimmer (Electromotive PC4215)
C502A,C502C	1000 pF, 5%, silver mica
C503A,C502B, C503C	1220 pF, 5% silver mica (1000 and 220 pF in parallel)
C503B	1500 pF, 5% silver mica
C504,C505	47 pF, 5% silver mica
C506	220 pF (see text)

capacitor parallels C105 only with the bandswitch in position C, and is necessary to increase the coupling and compensate for the large shunt capacitance required to hit resonance at 1.8 MHz. This completes the modification required to permit the rf tuned circuits to cover the range of approximately 1.8 MHz to 1.9 across C130 (switch section C). Also, as in *section 1,* install a 47-pF silver-mica between terminal 5 of S101B and terminal 5 of S101C. This completes the required rework except for alignment.

Before continuing with the alignment it's a good idea to be sure no loose bits of wire or solder are inside the receiver, and

that all other bands still work properly. Make necessary repairs by correcting any wiring errors, etc.

alignment

The *calibrate* function of the receiver may be used in lieu of a signal generator. Just remember to keep the rf gain control set so the S-Meter reads somewhere between S-5 and S-9. Final touch-up may best be performed "on-the-air." The reciver agc switch should be set to *fast agc.*

Set the preselector tuning knob to *9* and the main tuning dial to *0*. Set the bandswitch to C; this should represent 1.8 MHz. Turning the *calibrate* switch on and off should result in a signal easily identified. Leave the *calibrate* switch on, and by using the S-meter adjust the three trimmers just installed in *section 1* for maximum reading. Repeat the adjustment two or three times to be sure you have the correct adjustment. Move the tuning dial to *100*, which represents 1.9 MHz, turn the preselector knob to *1*, and noting the positions of each of the three trimmers carefully touch up the adjustments and again note the position of the trimmers. There should be very little difference between the two settings. Leave the trimmers set to the center of the two settings.

If you do not hear the 100-kHz marker tone at *100* on the dial, your crystal may not be oscillating on the proper frequency. Recheck the wiring of the rear section of S101 and readjust the trimmer capacitor (C127). This completes the adjustment of the 1.8-1.9 MHz segment of the 160-meter band.

The bandswitch should now be placed in position B and the main tuning dial to *200*, which represents 2.0 MHz, and the preselector tuning knob to *1*. Again, you should hear the *calibrate* marker signal, and it should sound clean – no hum or roughness. Peak trimmer capacitors C120, C125 and C130 for maximum S-Meter reading, repeating two or three times to be sure of the best settings and taking into account the precautions mentioned previously regarding overloading the receiver. Set the main tuning dial to *100* again, set the preselector knob to *9* and follow the procedure mentioned above to optimize the trimmer settings.

If desired, you may simply peak up the trimmers "on-the-air" for those portions of the band which you prefer, although very little difference in performance has been noted on the several receivers with which I have had experience. It should be remembered that moving the bandswitch from position C to position B will not affect the incoming signal tuning, as far as the *frequency* is concerned, but only from the viewpoint sensitivity. Extremely strong signals will be heard even though the preselector is not tuned to resonance with the incoming signal, but weak signals will be lost. There will, of course, be some overlap between the two segments near the 1.9-MHz portion of the band.

conclusion

This modification results in a receiver capable of good performance on 160 meters, and will provide all the sensitivity that can normally be used in most locations. Some operators may feel that an external preamplifier might be useful, but this is true only if the receiving location is extremely quiet and atmospheric noise is very low.

The sensitivity of the receiver I modified measured better than a half microvolt for a 10-dB signal-plus-noise-to-noise ratio. This was on both the CW and ssb modes of operation. This sensitivity should be sufficient for even serious DXing on 160 meters.

With a little study and a few easily obtainable parts this entire project can easily be completed in one evening's time and adds greatly to the usefulness and the pleasure of using the HQ215. I would like to thank George Faatz, K4KH, who proofread these instructions while performing the modification on his own HQ215.

ham radio

compact tone-burst keyer for fm repeaters

This simple tone-burst oscillator circuit provides high performance in a miniature package

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As two-meter fm repeaters multiply across the country, more of the groups operating them may be expected to turn to some form of tone-guarded input, not to exclude non-members, but simply to eliminate accidental sporadic keying of the system by stations using the same input frequency in another area.

With the decision to install a 2000-Hz single-tone decoder on the 146.34 MHz input to the WB8CQR repeater this past spring, I started looking for a replacement for the tone generator I built a year ago for my occasional visits to Chicago and other areas with controlled-access repeaters. That unit, a Colpitts-type transistor oscillator using an 88-mH toroid, keyed manually with a push button, worked perfectly but was bulky and awkward.

What I sought was an oscillator that would: key automatically with the push-to-talk circuit, provide a frequency-stable tone on at least two switch-selected frequencies, be small enough to mount inside one of the compact transistorized transceivers, such as my Varitronics IC-2F, and draw minimum current to avoid undue drain on a portable's power supply. The answer was found in a Motorola application bulletin.[1] It is a saw-tooth generator using two diodes and a capacitor, switched by a modified monostable multivibrator which can be triggered by the microphone push-to-talk circuit. The circuit is shown in fig. 1.

circuit operation

Operation is as follows: When the PTT switch is open (the receive condition) the microphone PTT circuit is ungrounded and a positive voltage appears at terminal 3. Applied to the base of Q1 through diodes CR4 and CR3, this voltage turns Q1 off, which turns Q2 on. It also charges C1.

With Q2 on, approximately 12 volts is applied to CR2, a Motorola field-effect diode (also known as a current-limiting diode or current Zener). Basically this device is a field-effect transistor with the control gate and source internally connected to the substrate; it has the interesting characteristic that when reverse biased it will conduct an almost constant current regardless of the applied voltage. This current ranges from 0.22 mA for the 1N5283 to 1.60 mA for the 1N5303; the circuit shown uses a 1N5299, rated at 1.20 mA ±10%.

The constant current through CR2 charges the capacitors at a constant rate, increasing the voltage across them and diode CR1. Diode CR1 is a four-layer or pnpn diode. It has two stable states: off, in which it exhibits resistance of a megohm or more, and on, in which its resistance is only a few ohms. Transition from off to on is controlled by the voltage applied; as the capacitors charge the diode remains off until its breakover voltage is reached, at which time it turns on, discharges the capacitors practically instantaneously and turns off again. The frequency of oscillation is thus determined by three factors: The current conducted by CR2, the value of the capacitors and the breakover voltage of CR1. A vernier adjustment is obtained by connecting a resistor across the capacitors to bleed off part of the charging current.

The oscillator runs constantly as long as the PTT switch is open and positive voltage is present at terminal 3. When the transmitter is keyed by closing the PTT switch and grounding terminal 3, the oscillator continues to run, and a tone is transmitted until capacitor C1 discharges.

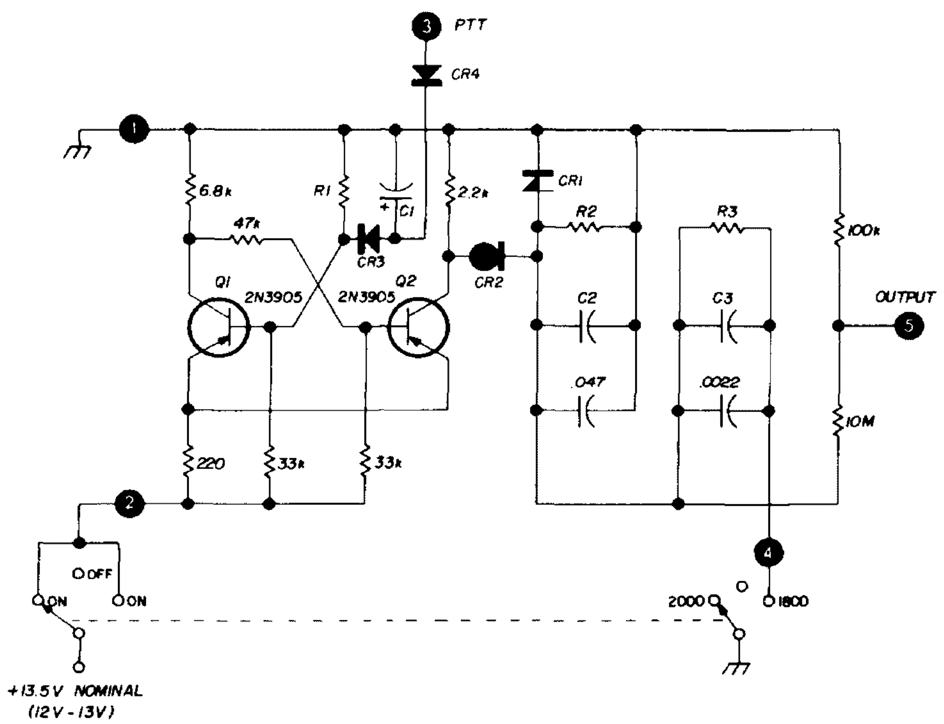

fig. 1. Simple tone-burst oscillator circuit provides tones of 1800 and 2000 Hz. Diodes CR3 and CR4 are general-purpose diodes. R1 and C1 are selected for desired burst duration; for value of C2, C3, R2 and R3 see text.

C1 and R1 thus control the burst duration; values of 30 μF and 100k ohms give about ¾ second. When C1 is discharged, Q1 turns on, Q2 is turned off, the voltage at CR2 is reduced to approximately 1 volt. The oscillator stops until the positive voltage is restored to the base of Q1 by release of the PTT switch.

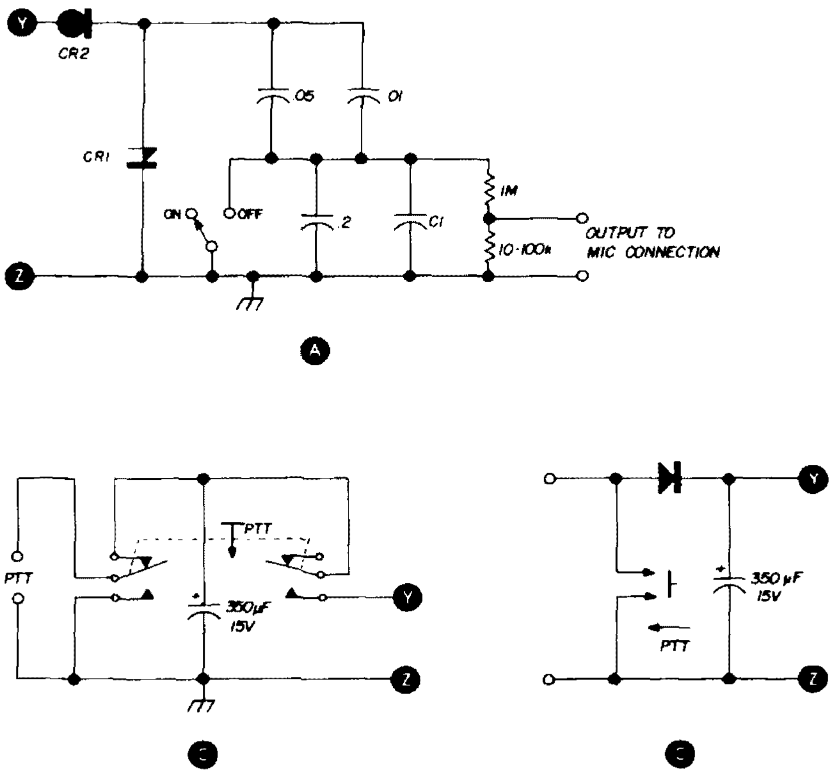

fig. 3. Most simple tone-burst oscillator can be built into a microphone. The circuit is powered by a charged 350-μF capacitor; as the capacitor discharges, the tone drifts downward. Capacitor C1 is approximately 0.02 μF (select to desired frequency).

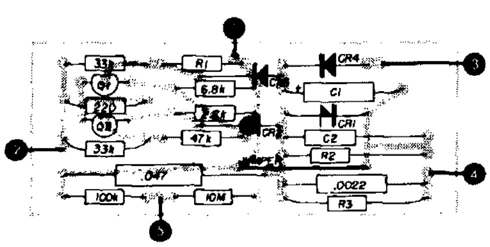

fig. 2. Full-size circuit-board layout for the compact tone-burst oscillator.

The unit draws approximately 6 mA with the oscillator running (1.2 mA through CR2 and the rest through the 2.2k load resistor connected to Q2). It draws less than 2 mA when the PTT switch is closed. It is very compact; the actual size of the circuit board shown in fig. 2 is ¾ by 1¾ inch, and the clearance required above the board is only 3/8 inch.*

construction

For most compact assembly all resistors for which numerical values are shown should be installed first, as close to the board as possible. Capacitors fit between and overlap the resistors. Transistors and diodes are installed in a second layer, with leads extending down to the board between the other components.

With all components in place except C2, C3, R2 and R3, the burst duration should be checked by observing the voltage at the collector of Q2 as the PTT

*Printed-circuit boards are available from the author for $2.50, postpaid, in the United States and Canada.

switch is closed. Adjust R1 if necessary. The oscillator frequency adjustment must be made experimentally using a counter or other standard, as the capacitors, diodes and resistors all have 10% tolerance and the allowable margin of error in most repeater tone decoders is on the order of 0.5%. Select C2 and R2 to give the desired high tone (2000 Hz in the unit shown here); then select R3 to give the low note. Installation of C3 is usually unnecessary.

The circuit has been tested in a variety of equipment with the WB8CQR single-tone decoder and keys the repeater with less than 2-kHz deviation on the transmitted signal. Deviation can be set at whatever level is desired by changing resistor values in the output voltage divider. Frequency is comparatively stable for a power supply voltage range between 12.5 and 15 volts. There is some drift with temperature change; if it is excessive, try new resistors and capacitors in the oscillator. I have found a few that were much worse than average and resulted in unsatisfactory performance of the unit in which they were installed.

The same oscillator circuit was used to make a miniature tone generator that could be completely contained in the mobile microphone case. The circuit is shown in fig. 3. Power for the oscillator is drawn from a 350-μF electrolytic capacitor charged from the PTT circuit. It had only one fault; the tone tailed off in frequency as the capacitor discharged, and the signal swept from 2000 Hz to below 1800 Hz before the oscillator cut off. This circuit is also sensitive to lead lengths in the capacitor network. C1 must be selected in its working location. If you want a universal repeater keyer, this will do it! If you value the opinion of your friends, however, I recommend the version shown in fig. 1.

reference

1. John Bliss and David Zinder, "Four-Layer and Current-Limiter Diodes Reduce Circuit Cost and Complexity," Motorola Applications Note AN-221.

ham radio

cw break-in circuit

Fast break-in is a must for traffic nets and contests — here's a circuit using solid-state switching for quiet operation

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Several simple break-in circuits for CW have been tried at my station with varying degrees of success. Some seemed to work fairly well, but key clicks and thumps were difficult to eliminate. Mechanical relays for switching increased the room noise level considerably. When the break-in system is too noisy, one tends to send faster or shorten character spacing to hold the relays closed, thus reducing noise. This, of course, defeats the purpose of break-in, which is to allow signals to be heard between letters while sending. Other circuits I have not yet tried seemed far too complicated for simple break-in.

preliminary steps

Before attempting to use any break-in system, the receiver and transmitter should be tested to determine if they will perform properly under fast switching conditions. In full break-in, the exciter and final-amplifier plate voltages are on and the tubes biased to cut-off or beyond. This condition could cause diode noise in the receiver and should be corrected.* The receiver should be tested for its willingness to cooperate with the break-in system by connecting a variable voltage, 0-75 volts, to the avc line. With bias connected key the transmitter. If a clean signal is heard without key clicks or thumps and the gain can be adjusted to the desired level, full break-in circuitry can be added with no difficulty.

break-in circuit

A separate receiving antenna is used at my station. In the circuit shown in fig. 1 a switch, SW1, is used to change modes. The circuit is designed to be used with grid-block keying, and the −75 V is obtained from the transmitter bias supply. A separate supply could be used, however.

*Although this problem can be caused by tube-type rectifiers in the transmitter high-voltage supply (due to contact potential, leakage paths, etc), a common cause of receiver noise under the conditions stated by the author is improper bias and neutralization of the final amplifier tubes. **editor.**

When the four-pole, three-position switch is in the break-in position, SW1D connects the receiver to the receiving antenna, and the transmitting antenna input is grounded at this point through SW1C. Capacitors C3, C4 and diode D2 form a solid-state switch, which is quite effective. This circuit is an improvement over simply shorting the antenna to ground and is more effective than a spdt relay.

SW1B connects the key or electronic keyer to the base of Q1, an npn silicon transistor. With −75 V connected to the emitter and the base resistor, R1, returned to ground through the key, Q1 conducts heavily, and the −75 V appears at the collector and at the top of R3. Since R3 is connected in series with D2, D2 also conducts heavily, grounding the antenna input to the receiver. A variable bias voltage of 0.6 to −75 V is now available at the slider arm of R3 and is connected to the receiver avc line through D1. D1 prevents shorting the avc line during key-up conditions.

SW1A turns the transmitter B+ on in both break-in and normal transmit positions. In the transmit position, Q1 conducts and the receiver may be used for monitoring but will remain off until the switch is returned to receive or break-in. In the receive position, the receiver operates with the transmitting antenna, but the solid-state switch is still in the circuit.

construction

The entire circuit is built in a 2 x 3 x 5-inch minibox. Phonograph jacks and plugs are used for all connections. Shielded leads are used. The transistor I used is a Motorola RV video amplifier type M4843. Television receivers, especially color sets, use transistors with rather high voltage ratings. Some horizontal sweep power transistors operate with peak voltages as high as 800 volts. In this circuit any good quality npn transistor of the required voltage rating may be used, but it would be a good idea to check for leakage before closing the box. No voltage should be present on the collector with the key open in the receive or break-in switch position.

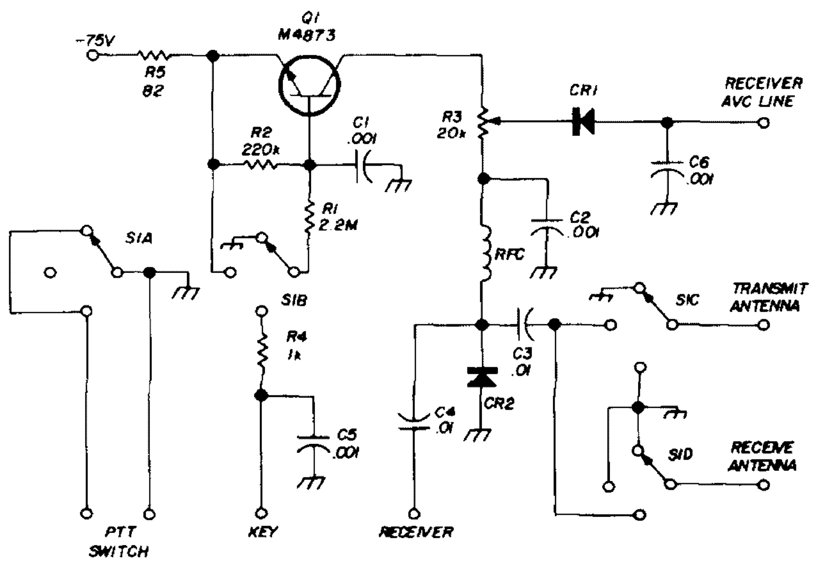

fig. 1. CW break-in circuit using solid-state switch for stations with separate transmitting and receiving antennas. Design is for grid-block keying systems.

I've been using the circuit for a few months and have been pleased with the results. Received signals can be heard between letters at reasonable code speeds. I use the circuit with an unmodified 75A4 receiver on fast avc. The receiving antenna consists of about 25 feet of wire and works quite well except for signals below S4 or S5. A better receiving antenna will be installed soon, which should improve circuit performance.

ham radio

low-pass filters

for 10 and 15 meters

Data for building two low-pass filters with better harmonic attenuation than conventional designs

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Low-pass filters have been designed and built for many years according to the conventional formulas found in various handbooks. Such formulas are based on so-called image-reflection design, and a filter built according to these formulas consists of one to three constant-k center sections and two m-derived terminating half sections. The filter is usually designed for use with a 52- or 72-ohm coaxial line. The attenuation characteristics of such a filter are valid only if the filter is used with the type of line for which it is designed and if that line operates near unity swr. In recent years, filter design has improved because of an interest on the part of military and other communications-equipment designers to achieve superior filter performance at minimum cost.

This article presents construction data on low-pass filters for 10 and 15 meters which have been designed according to latest available techniques. Either filter may be built from readily available parts for any power to 2 kW PEP.

The advantage of filters designed according to modern techniques is the superior harmonic attenuation obtained using fewer components. For example, the filters described here use only 3 coils; yet the harmonic attenuation is as good or better than that of filters using 5-6 coils. Using fewer coils may not sound like an advantage in theory, but it means a great deal in practice. Unlike capacitors, coils in a low-pass filter must either be shielded or oriented to prevent mutual coupling, which degrades filter performance.

Building a really effective low-pass filter that gives a *true* 50-70 dB harmonic attenuation, using conventional design techniques, often presents a mechanical problem because of the shielding required between filter elements. Modern filters, on the other hand, are not only easier to build to achieve the same performance level but are less critical as to construction techniques. The latter assertion can't be proved from theory, but it certainly seems to be the case as far as I've been able to observe when building various filters.

design characteristics

The filter design is shown in **fig. 1**. It's

intended for insertion in a coaxial transmission line of 52-72 ohms (design optimized for 52 ohms), either between an exciter and a linear amplifier or between a linear amplifier and antenna. The three-section design uses a combination of parallel- and series-resonant circuits. The design can't be reduced to only two coils, because all circuit elements are interdependent. Built as shown, however, the filter yields outstanding harmonic attenuation without involved construction details.

The filter design has a cut-off frequency up to which the filter will pass radio frequencies with minimum attenuation. Beyond the cut-off frequency, the attenuation increases rapidly with frequency until it reaches the first peak of maximum attenuation. In the specific designs shown later, this latter frequency is from 1.5 to 2 times that of the cut-off frequency. At this first attenuation peak, the attenuation is theoretically at least 70-90 dB as compared to the level below the cut-off frequency. The attenuation at all higher frequencies is at least 70 dB, with a number of peaks providing even higher attenuation.

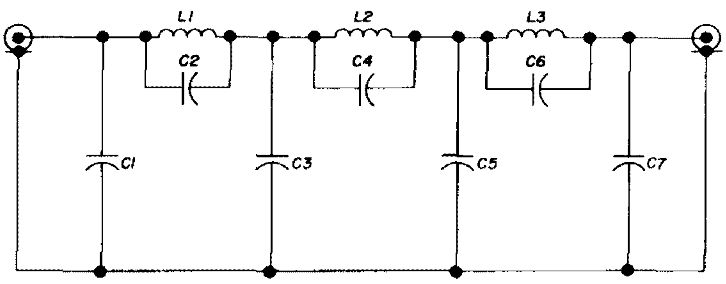


fig. 1. General arrangement of filter. Refer to table 1 for constants.

practical considerations

One can design filters that achieve theoretically even greater harmonic attenuation. However, the practical realization of levels greater than 70-90 dB with simple construction materials and techniques is illusory. The practical limit for filter elements seems to be about 120 dB, and this has been achieved only in designs using the most elaborate and expensive construction techniques. A paper design may achieve almost an infinite harmonic attenuation level, and, indeed, this is one of the dangers of conventional filter design where it appears that one could add constant-k sections almost endlessly. In short, the design shown in fig. 1 represents the practical limit for a filter that will work well.

15-meter model

The first filter to be described is for use with transmitters operating at 21 MHz or below. Component values are shown in table 1. All may be achieved with readily available capacitor values. The 15-meter filter attenuation begins about 22-23 MHz and reaches its first peak (about 60 dB) at 40-42 MHz (**fig. 2**), which provides good harmonic attenuation in the i-f range of most TV sets. This feature is, in fact, the reason for developing a 15-meter filter. A low-pass filter that begins cutoff above 21 MHz provides excellent attenuation at TV intermediate frequencies as well as at the beginning of the vhf TV channels. Another consideration is that, as the sunspot count decreases in the coming years, there will be less activity on 10 meters.

Many cases of TVI can be eliminated by this means where TV i-f interference has been a problem. In a few instances, even problems with nearby CB installations can be avoided with such a filter since its attenuation at frequencies above the 21-MHz band will be effective for both reception as well as transmission.

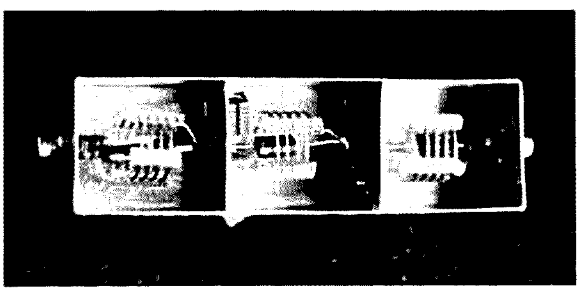
10-meter model

The second filter has a cut-off frequency of about 30 MHz. An attempt was made to design a filter that offers significant protection at TV-receiver i-fs using filter elements within reasonable tolerances. Circuit values for this filter are also shown in table 1. The first maximum attenuation peak occurs about 60 MHz. However, the slope after the cut-off frequency is quite steep, and significant attenuation begins at 40 MHz. Again, the filter was designed for available capacitor values.

capacitors

Mica capacitors should be used. The physically smallest 500-vdcw mica is inexpensive and may be used for filters operating at the 300-watt PEP output level (15 meters) and at 100-150 watts for 10 meters.

These capacitors therefore, are suitable for a filter between a 100-watt (nominal)



Typical filter using brass sheet stock. This design was intended for use between an exciter and a linear amplifier. Receptacles are rf-type phono connectors.

output exciter and a high-power linear. Such a filter is often overlooked in TVI problems because of the expense of a commercial filter for both the exciter-final link and the final-antenna link. However, a filter between exciter and final is useful in attenuating harmonic radiation, which causes TVI, because the exciter will have some harmonic output.

Larger mica capacitors may be used for filters for any desired power level. Molded mica capacitors rated at 1500 volts dc working are useful in filters operating to 1-kW PEP *output*. Type CM-15 mica capacitors, rated at 2500 or

table 1. Practical component values. Capacitor values are based on commercially available parts. Try not to substitute.

component C in pF L in μH	15-meter filter (22 MHz cutoff)	10-meter filter (30 MHz cutoff)
C1	68	50
C2	7	5
C3	200	100 & 39 parallel
C4	27	20
C5	180	100 & 33 parallel
C6	22	15
C7	57	47
L1	.41	.30
L2	.45	.33
L3	.35	.26

5000 vdcw, may be used with even the most efficient 2-kW PEP linear.

construction

Filter construction is easy since only three coils are used, which should be mutually shielded. The coils can be made from Air-Dux or B&W stock, or wound from tinned copper wire. The small sizes of coil stock are fine for transmitter *output* powers up to 500 watts PEP. Coils for higher power should be wound with no. 10 or 12 tinned copper wire. The capacitor values in table **1** should be adhered to.

The photo shows a filter using low-voltage mica capacitors. Brass sheet stock forms an enclosure 5½ inches long by 2 inches square. Equally spaced dividers separate filter sections, and simple feed-through insulators are used between sections.

Brass sheet has several advantages. The material is easy to work. The sides of the enclosure as well as the dividers can be soldered. The cover can also be completely solder-sealed. Such shielding efforts really help, particularly when trying to combat TVI in the uhf ranges where the screwdriver assembly of conventional filters allows rf leakage.

results

The attenuation characteristics of the filters I built are shown in fig. **2**. The 15-meter filter slope reaches an attenua-

tion peak at the TV i-f range. The equipment I used to check the filter was of good laboratory quality, which the average amateur doesn't have. Nonetheless, the instrumentation wasn't capable of making *accurate* attenuation measurements in the vhf range beyond -60 to -70 dB. The curves of **fig. 2**, therefore, are conservative.

A careful sweep of the vhf range disclosed no frequency where filter attenuation fell below -60 to -70 dB. This result included the effects of the phono-type connectors used in the filter construction. The attenuation of the filter in the pass-band region varied between .1 and about .25 dB but never exceeded the latter value. When inserted in a matched 52-ohm transmission line, the filter produced a worstcase swr of 1.1.

conclusion

I claim no credit for the original design of the filter. That belongs to a number of scientists. The purpose of this article is to present two selected and truly practical low-pass filters that satisfy requirements for any 80-10 meter amateur installation.

A point that should again be emphasized is that the filters *must* include all the sections to be effective. You can't eliminate any section without completely changing the filter's attenuation characteristic.

ham radio

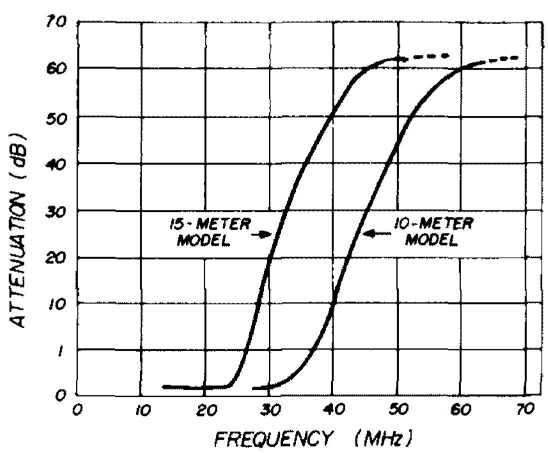


fig. 2. Measured results of filter attenuation characteristics.

threshold-gate/limiter for cw reception

Solid-state diodes in a simple and effective circuit to reduce filter ringing

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A disadvantage of audio filters and Q multipliers is their ringing effect on received CW signals. This ringing is caused by the shaping effect introduced in the audio signal by the filter or Q multiplier, namely, rounding and lengthening, respectively, of the signal leading and trailing edge. A method of reducing this distortion is to introduce a later distortion that sharpens and shortens the signal.

threshold gating

A characteristic of solid-state diodes, i. e., their ability of not conducting in the forward direction until a critical voltage is reached, can be used to decrease ringing and background noise as well. Fig. 1 shows how a pair of diodes can be used to counteract ringing by providing a threshold-gating effect. The circuit has two additional features – it eliminates background noise below threshold level and provides single-signal reception of all signals with audio images falling below threshold. Further sharpening can be accomplished with a second set of diodes having a higher knee voltage, which operate as limiters. The effect of the combina-

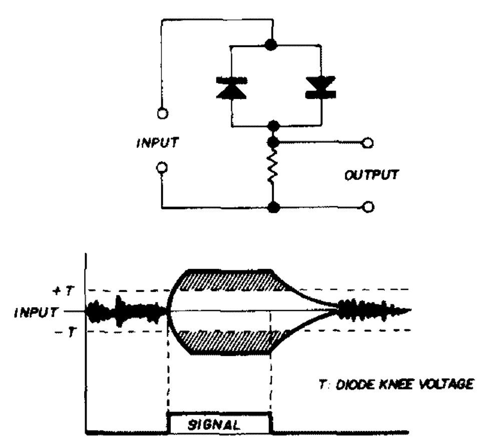

fig. 1. Threshold-gating. Input waveform represents ringing effect of high-Q filtering; shaded parts represent output of threshold-gating circuit. Note elimination of background noise and reduction of ringing.

tion is shown in fig. 2 and can be described as a window-effect. The limiting diodes also prevent blasting by loud signals and, as shown in fig. 2, materially reduce the effects of signal fading.

A criticism of the threshold-gating/limiting circuit is that it severely distorts any audio tone passing through its intensity window. This is a valid criticism, and such distortion occurs when the circuit is used without prior high-Q filtering – the tone sounds raspy or scratchy. However, with the introduction of high-Q filtering the tone changes to a rather pleasant sound. Best results were observed when both Q multiplier and audio filter were used. It can be readily demonstrated that high-Q filtering and threshold-gating/limiting are effective in combination.

practical circuit

A practical circuit appears in fig. 3. Diodes may be selected from commercially available devices. In this case, I found that the emitter-base junction of a 2N414 made a good threshold diode, while the 1N54A performed well as a limiter. A switch was included so that the threshold-gating/limiting effect could be disabled for receiver tuning and when a signal quickly drops below threshold. A foot switch was convenient for this function. Note that the receiver output is not properly terminated. The problem here is

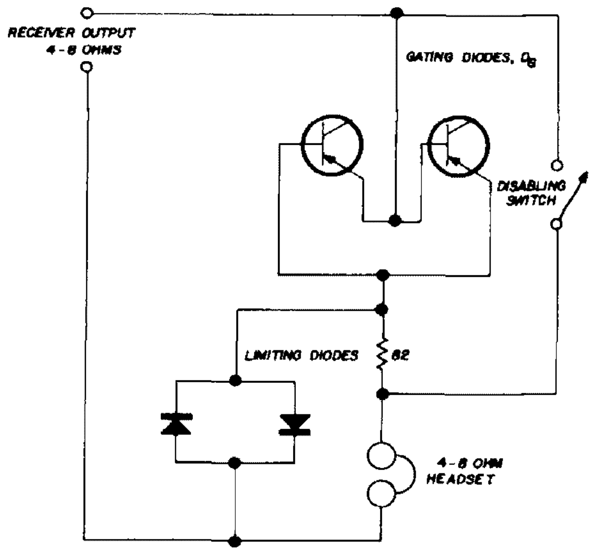

fig. 3. Intensity window practical circuit. Emitter-base junction of a 2N414 is used for the gating diode and 1N54A as limiting diode.

the possibility of abnormally high voltages in the receiver output circuit causing damage to components. However, the limiting diodes reduce this possibility by clamping the output when high-level signals are received. The circuit has been used with vacuum-tube receivers without adverse effect. Although low-impedance earphones are indicated, high-impedance earphones may be used. These may be substituted directly, but it's preferable to use a matching transformer such as a vacuum-tube audio output transformer.

With high-Q filtering, the circuit provides the unique experience of being able to tune across a CW band and hear signals of approximately equal amplitude and tone drop in and out of a noiseless background. All signals tend to sound alike in terms of keying characteristics, as the combination of high-Q filtering and threshold-gating/limiting tends to eliminate individual differences. Most signals seem to be improved by the process. The circuit can be built into the cord of a headset.

ham radio

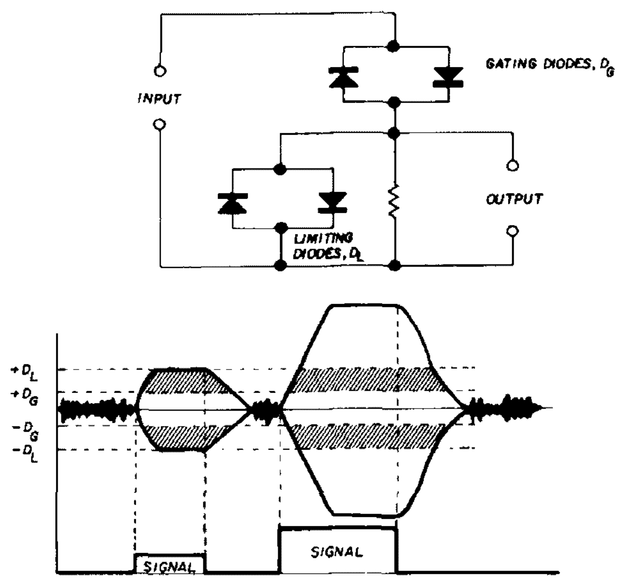

fig. 2. The intensity window. The combination of threshold-gating and signal limiting restrict output (shaded region) to components falling between the knee voltages of the two sets of diodes. Note further sharpening of input signal and reduction of fading effects.

rf bypassing at vhf

Suggestions for choosing the correct bypass capacitor in vhf circuits on PC boards

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If you're planning to build equipment for use at vhf, the following suggestions will be helpful in selecting values of small capacitors for rf bypassing.

Above the self-resonant frequency of a capacitor an inductive component appears that can degrade the effectiveness of the capacitor as an rf-bypassing element. At frequencies near 100 MHz the problem is compounded when equipment is built on PC boards.

capacitor frequency characteristics

The inductive reactance that predominates in a capacitor above its self-resonant frequency is often ignored by many amateurs. To them, a capacitor provides ac coupling, ac bypassing, and dc blocking. Many also believe that the larger the capacitor, the better it will bypass high frequencies. The fallacy in these assumptions will become apparent in the following discussion.

Consider fig. 1. At A is a "perfect" capacitor. It has zero lead length, zero electrode dimensions, and a lossless dieletric. However, when plates are added to the "perfect" capacitor, and leads are added to the plates, a certain amount of inductance is introduced, as shown in B. The inductance in the wire leads is approximately 20 nH (0.02 μH) per inch of lead length.

Now suppose the capacitor of fig. 1B is used in a bypass circuit on a PC board; 1/16-inch PC-board foil has an inductance of 20 nH per inch. If we try to bypass the emitter of a 100-MHz transistor amplifier with a large-value capacitor having ½-inch leads, and the PC-board foil between bypass capacitor and emitter lead is ½ inch, a total of 1½ inches of lead length will exist between the transistor emitter and ground. At 20 nH per inch, this results in 30 nH between emitter and ground. At 100 MHz, the inductive reactance of this combination is of the order of 20 ohms.

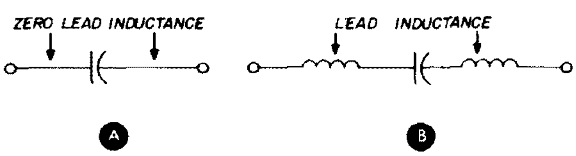


fig. 1. A "perfect" capacitor has no inductance. When leads are added, their inductance degrades the capacitor's effectiveness as a bypass element at vhf.

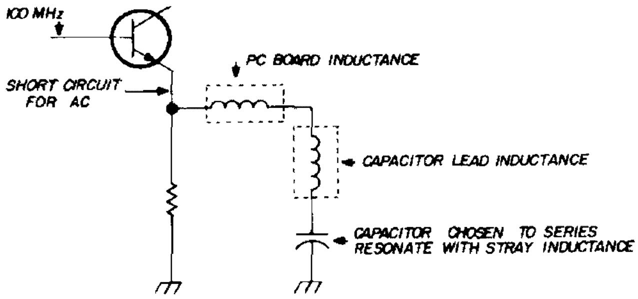


fig. 2. Effect of added inductance to a vhf-bypass circuit by PC-board foil and capacitor leads.

reactance considerations

The idea is to choose a capacitor that is series resonant at the desired bypass frequency. For instance, in the above example, a 68-pF capacitor with ½-inch leads and ½ inch of PC-board foil will series resonate at 100 MHz. The circuit will have minimum impedance at this frequency, and maximum current will flow. Thus the circuit will provide an effective by-pass for unwanted rf energy to ground. (See fig. 2.)

In summary, at frequencies above a capacitor's self-resonant frequency inductive reactance predominates, and the capacitor doesn't function efficiently as a bypass element. Below its self-resonant frequency, capacitive reactance *predominates* and the capacitor looks capacitive to the circuit. This relationship is shown in fig. 3.

graphical aid

Fig. 4 shows experimental results measured with small disc capacitors and a grid-dip oscillator to find resonant frequencies of various capacitor values.

To use the graph in designing a bypass circuit, first ensure that the lead length *between the point to be bypassed and* ground is as short as possible. Then choose a capacitor value that self-resonates at the frequency to be bypassed.

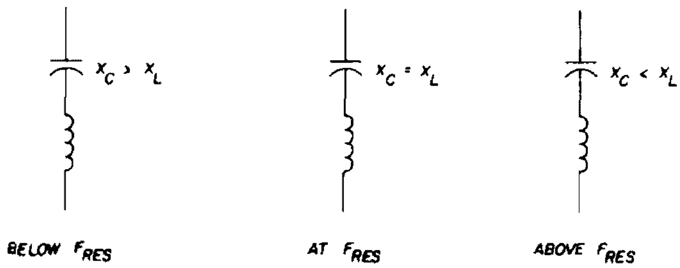


fig. 3. The behavior of capacitive reactance in a capacitor in terms of its self-resonant frequency.

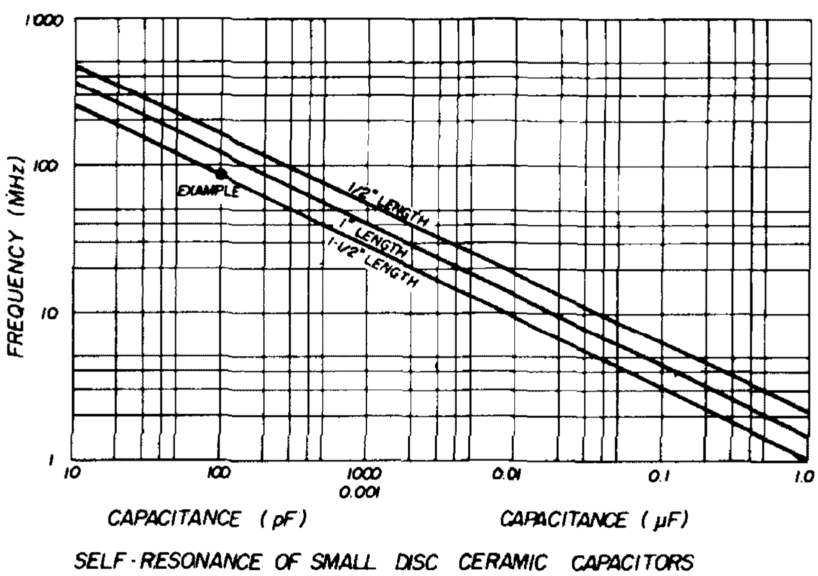
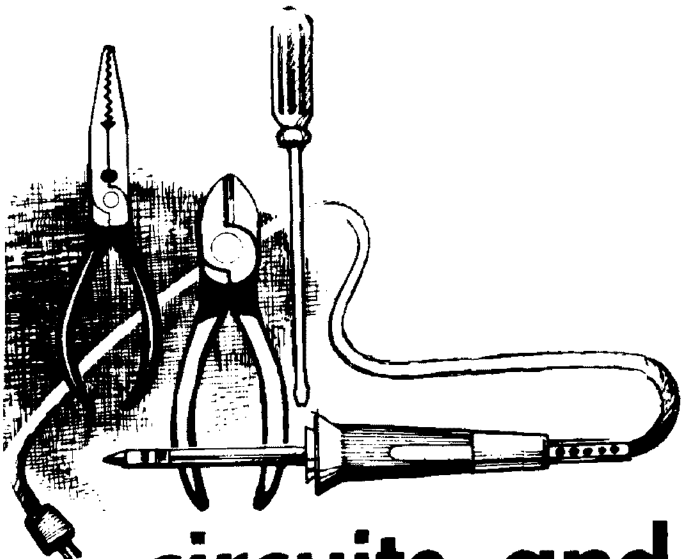


fig. 4. Aid for choosing disc capacitors in bypass circuits at vhf. All lead lengths between active elements and ground should be as short as possible.

In short (no pun intended), choose your bypass capacitor with care or you might as well not have a bypass at all.

ham radio



circuits and techniques

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amateur applications for wwv-wwvh

WWV (Fort Collins, Colorado)[1], and WWVH (Kauai, Hawaii) transmit frequency and time signals plus other data that provide a variety of helpful services in the modern ham shack, from setting the station clock to measuring modulation. Many frequencies provide calibration points for receivers, frequency meters and calibrators. Standard audio tones permit calibration of audio sources used for testing ssb and f-m transmitters. One-second time-marking signals are transmitted continuously by both stations.

WWV broadcasts on carrier frequencies of 2.5, 5, 10, 15, and 25 MHz. WWVH transmits on the same carrier frequencies with the exception of 25 MHz. These standard signals, available day and night, are heard in most parts of the earth. The stability is exceptional, and frequencies are held stable to better than two parts in 10^{11}. In fact, deviations at WWV are normally less than one part in 10^{12} from day to day. Propagation variables such as Doppler effect, when they occur, usually result in a greater fluctuation in the carrier frequencies. This is a result of the path and not the standard.

signal makeup

In addition to serving as a carrier frequency standard, the modulation data carried by the WWV-WWVH transmissions is substantial. The broadcast format is shown in **fig. 1**. Transmissions from both stations are identical except for a few minor variations that will be mentioned.

The sequence and content for most minutes of each hour are depicted in **fig. 2**. At 00-second of each minute there is a 0.8-second long 1000-hertz tone on WWV and 1200-hertz tone on WWVH. Just prior there is a voice announcement of Greenwich mean time (gmt). Following the time pulse there are continuous second ticks that continue until the voice announcement that comes before the next time tone. The tone frequency is either 500 or 600 hertz, alternating between these frequencies each minute. The 29th second tick (pulse) is omitted.

The time announcement period extends between 45 and 60 seconds. At WWV the first half of this period is silent except for second ticks. The gmt time announcement is made in a male voice during the second half of this period. The positions switch for WWVH; the announcement comes first in a female voice and the silent period comes second. In those areas where both signals are re-

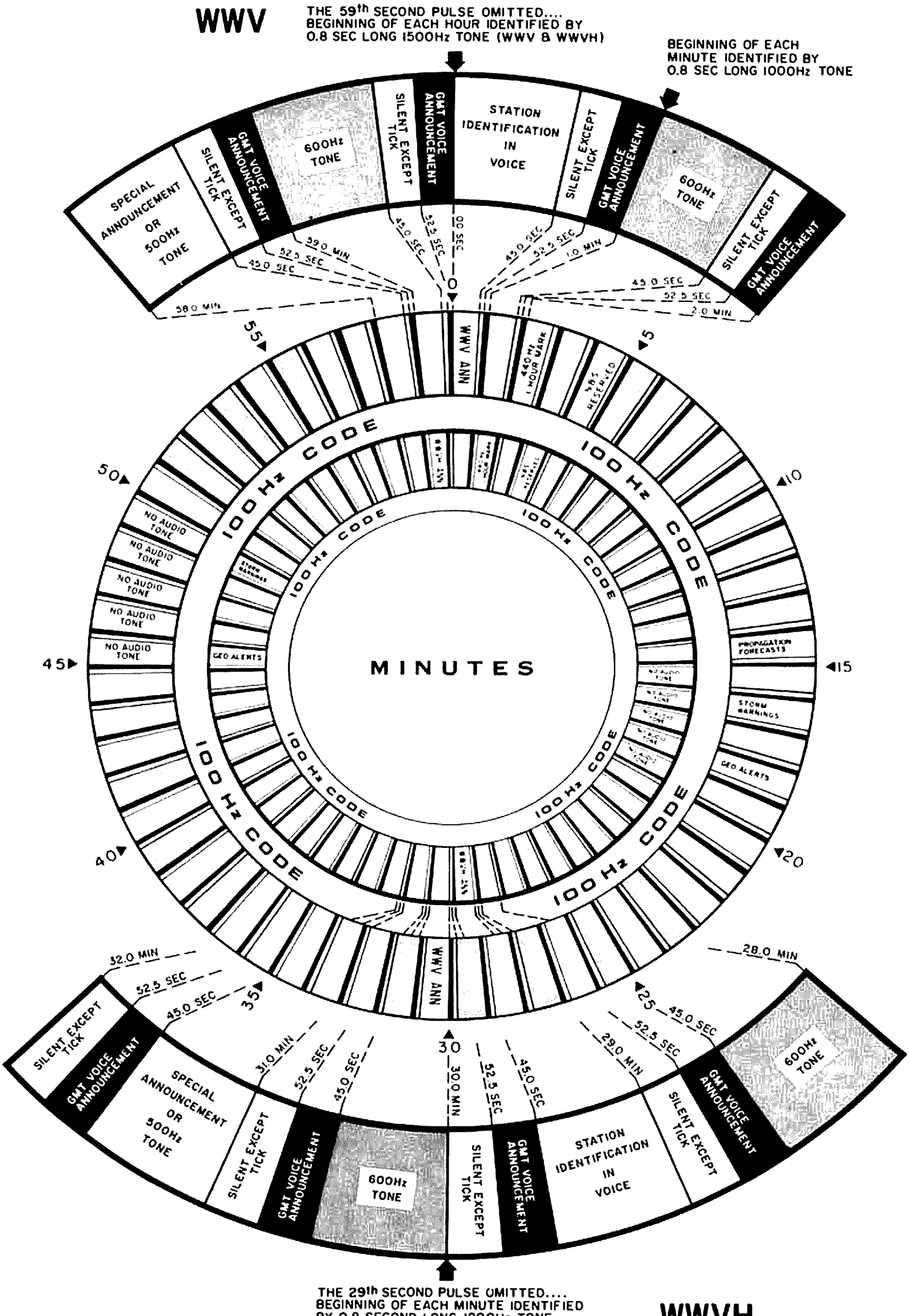


fig. 1. WWV and WWVH broadcast format.

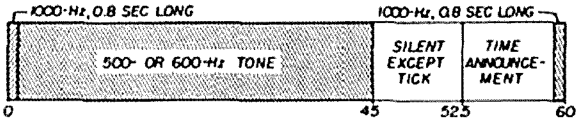


fig. 2. Typical content of a transmission minute from WWV. WWVH is similar except for 0.8-second segments.

ceived you hear the WWVH announcement first followed by the WWV announcement. The first minute of *each hour* begins with a 0.8-second, 1500-hertz tone on both stations.

Tone signals as displayed on an oscilloscope screen for 500 and 600 hertz frequncies are shown in **fig. 3.** This signal was demodulated in Pennsylvania from the 15-MHz carrier. Just above the 20-meter band this carrier is a good indicator of the transcontinental path quality on 20 and often for 15. When the WWVH signal is very audible in the background it indicates a path is open to the Pacific.

The stations identify on the hour and half hour. One station makes its announcement during the minute before the hour or half hour while the other station identifies itself during the minute after the hour or half hour.

Certain minutes of each hour are set aside for other emission sequences. As shown in **fig. 1**, no audio tone is transmitted between the 45th and 50th minute of each hour for WWV, and between the 15th and 20th minute for WWVH.

WWV gives propagation forecasts during the 14th minute of each hour, storm warnings during 49th minute.

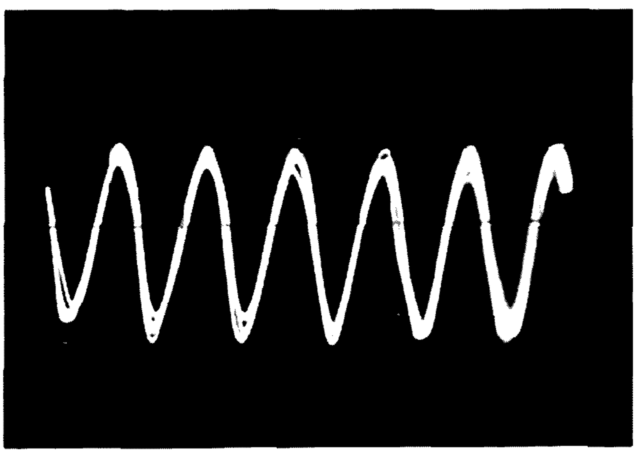
A 440-hertz tone is transmitted for minute. At WWVH the geophysical alerts are given during the 45th minute and storm warnings during 49th minute.

At 440-hertz tone is transmitted for approximately 45 seconds beginning one minute after the hour at WWVH and two minutes after the hour at WWV. This is a standard musical tone of note A above middle C. In addition, the tone can be used as hourly markers for various types of chart recorders and automated devices.

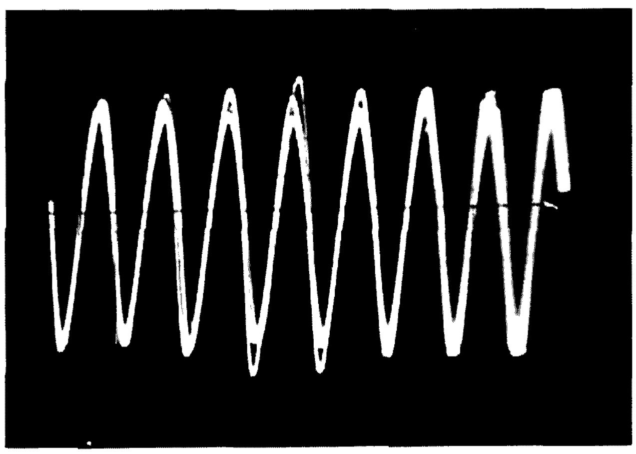
Special National Bureau of Standards transmissions are allocated a spot three minutes after the hour on WWVH and four minutes after the hour on WWV.

The makeup of the seconds-tick signal is shown in **fig. 4.** It occupies a time slot of 0.005 seconds and occurs once each second (one hertz repetition rate). The WWV pulse consists of five sine waves of 1000 hertz; the WWVH pulse, 6 cycles of 1200 hertz.

WWV and WWVH transmit a complex time code signal continuously. This code is produced at a one pulse per second rate and is carried on a 100 hertz subcarrier. It provides a standardized timing base containing time-of-year information based on universal time in seconds, minutes, hours and day of the year. It provides a standardized time for use when scientific



500 Hz pattern



600 Hz pattern

fig. 3. WWV tones as received and displayed on oscilloscope screen.

observations are made simultaneously at separate locations. Perhaps it may be a help for radio amateur satellite communications systems. Ten-millisecond resolution is obtainable from these transmissions.

audio calibration

The standard audio tones from WWV are excellent for calibrating sine- and square-wave audio oscillators. Fixed-frequency and switchable solid-state audio generators such as two-tone oscillators and audio generators for checking fm deviation can be constructed and set on precise frequencies using the WWV signal and oscilloscopic Lissajous patterns.

Tunable audio or square wave generators also can be set or checked on a number of frequencies using fundamental, harmonic or sub-harmonic patterns. The demodulated WWV signal is applied to the vertical input of the oscilloscope as shown in **fig. 5**. From **table 1** note the variety of frequencies that can be measured using the harmonics and subharmonics of the two standard tones. When necessary other components can be derived from the 440 hertz tone, although it is only transmitted for one 45-second interval each hour.

It would be no problem to construct a switchable tone generator with certain key frequencies that could check the audio filters of CW, sideband, fm and a-m equipment. Several subharmonic and

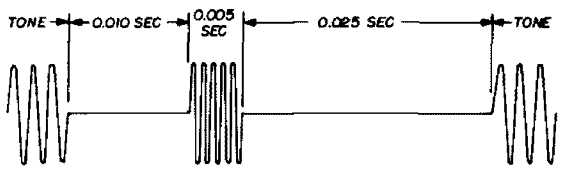


fig. 4. Makeup of second time pulses.

harmonic characteristic patterns are given in **fig. 6**. These were photographed at night while receiving the 10 MHz WWV signal.

stable audio oscillators

P. C. Lipoma[2] uses a Motorola

table 1. Checkpoints for calibration using WWV tones.

500 Hz WWV	600 Hz WWV	440 Hz WWV	Ratio
100	120	88	1-0.2
125	150	110	1-0.25
166+	200	146+	1-0.33
250	300	220	1-0.5
500	600	440	1-1
1000	1200	880	2-1
1500	1800	1320	3-1
2000	2400	1760	4-1
2500	3000	2200	5-1
3000	3600	2640	6-1

MC1454 integrated circuit in a Wein-bridge oscillator circuit, **fig. 7**. Frequency of operation is determined by the R1C1 and R2C2 time constants. Usually C1 and C2 are of equal value while the ratio of R1 and R2 is 2/1 to compensate for input and shunting impedances. Typically the capacitor value for 1000-hertz operation

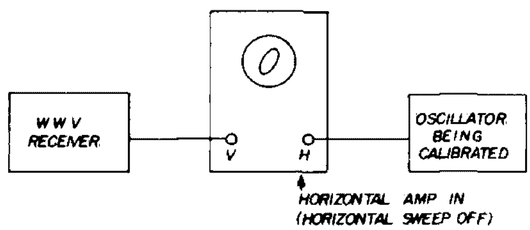


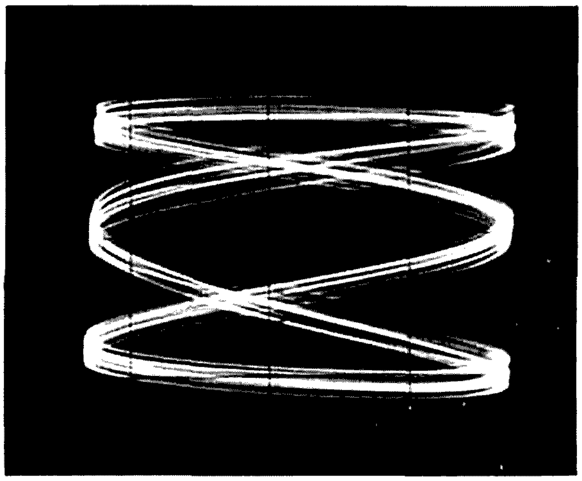
fig. 5. Setup for Lissajous calibration.

is 0.033 μF. To halve the frequency you would double the capacitor value; to double the frequency, halve the capacitor value. Resistor R2 could be made variable to permit precise frequency setting using the WWV standard tones. For example, a two-to-one Lissajous would set the oscillator precisely on 1000 hertz using the 500 hertz incoming standard.

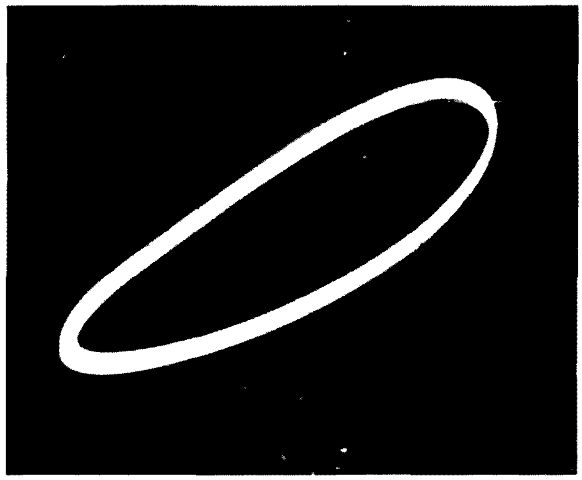
Two such oscillators would give you an excellent two-tone source. An oscillator with switchable capacitors would be good for checking audio filters.

radio-frequency calibration

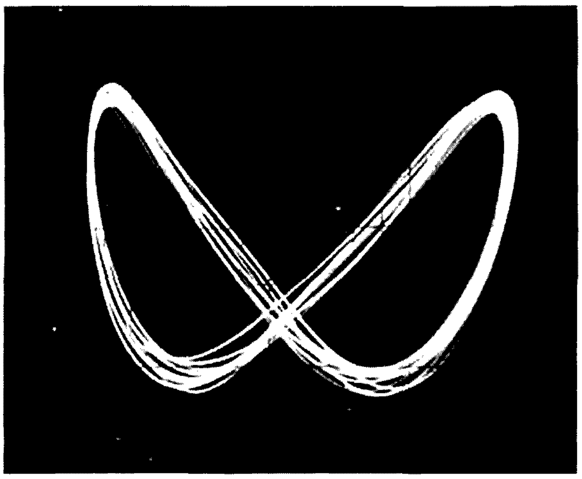
Radio-frequency calibration can also be made by applying the signal source along with the WWV signal to the receiver



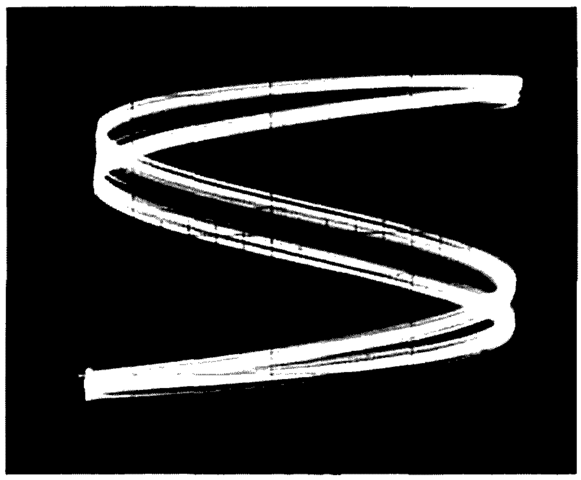
Ratio 1 to 0.33



Ratio 1 to 1

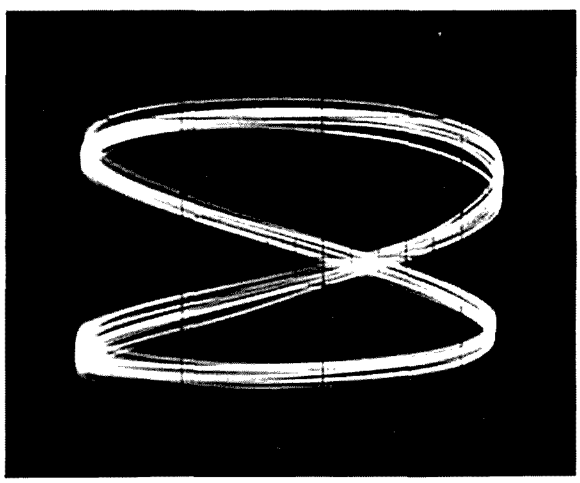


Ration 2 to 1



Ratio 3 to 1

fig. 6. Typical Lissajous patterns taken while receiving WWV. The test setup used is shown in figure 5.



Ratio 1 to 0.5

input, fig. 8. You can calibrate by ear or make a more precise check by using an oscilloscope connected to the receiver i-f system. The pattern swells on each side of zero-beat of the rf oscillator under calibration, fig. 9. You obtain a similar pattern when checking a 5-MHz or 7.5 MHz source as you bring its harmonic output to beat with the 15-MHz WWV signal. In fact, it is easy to calibrate a 100-kHz harmonic oscillator such as those used in the usual crystal calibrator with the same procedure. A good calibration time exists during the 45-50 minute no-tone time interval of WWV-WWVH transmissions.

propagation forecasts

WWV transmits propagation forecasts during the 14th minute of each hour. The telecommunications center of the Depart-

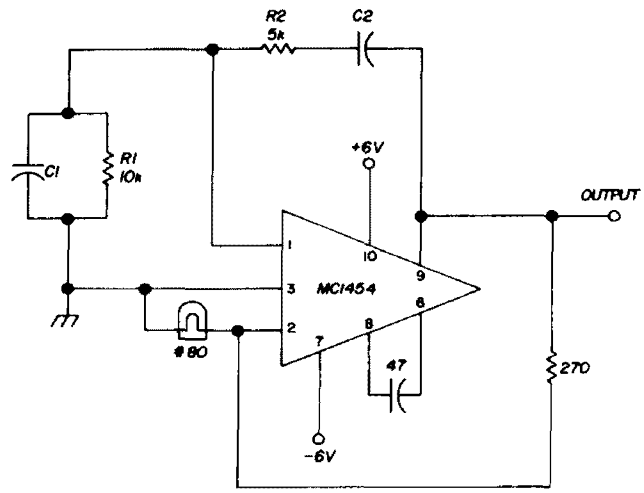


fig. 7. Integrated-circuit Wein-bridge audio oscillator. Making R2 variable would allow precise frequency adjustment using WWV as a standard.

ment of Commerce issues these forecasts daily at 0100, 0700, 1300 and 1900 gmt. Presently each forecast is broadcast unchanged on WWV until the next regular forecast is issued. They give an estimate of the radio quality expected on high-frequency radiocommunications circuits across the North Atlantic path. They apply to typical propagation paths such as Washington-Paris, New York-London and Washinton-Reykjavik. However, they are a good guide to propagation conditions for amateur band paths between eastern U. S. A. and Europe.

The forecast statement consists of a letter and a number. The letter is a state of current radio quality over the North

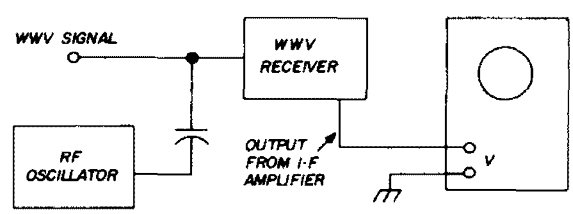


fig. 8. Setup for calibrating an rf oscillator with WWV signal.

Atlantic path at the time of issue. It is expressed in one of three grades:

N Normal or quiet
U Unsettled
W Disturbed

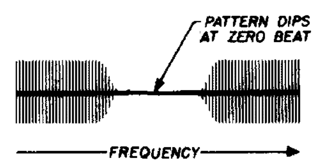
The number in the forecast states the expected value of radio propagation quality on a scale of 1 to 9:

1. Useless	5. Fair
2. Very poor	6. Fair to good
3. Poor	7. Good
4. Poor to Fair	8. Very good
9. Excellent	

geophysical alerts

Solar and geophysical information is broadcast on WWV 18 minutes after the hour and on WWVH at 40 minutes after the hour. Information is changed daily after 0400 gmt. However, outstanding events are added to the next hourly announcement as soon as they are identified. The information transmitted concerns solar activity, flux and flaring, proton events and flares and conditions of the geomagnetic field. For example the solar flux is announced and a forecast

fig. 9. Comparative oscilloscope patterns showing zero-beat.

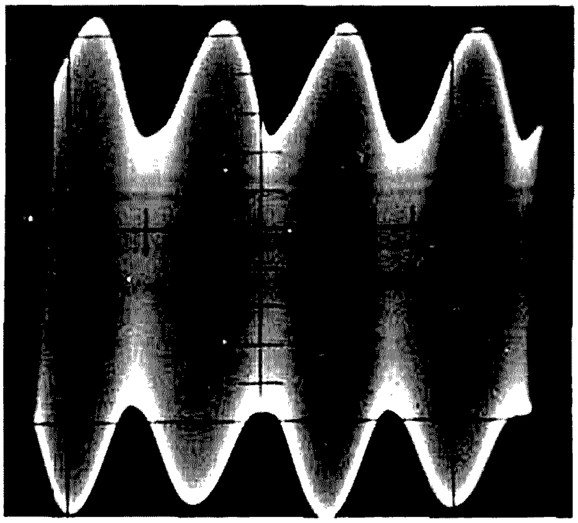


made whether solar activity is to be very low, low, moderate, high or very high. A geomagnetic field broadcast will state whether the field is to be quiet, unsettled or active and when a geomagnetic storm is to be expected.

Radio propagation conditions are affected by many factors including solar and geophysical events, and the weather. Hams who delve into the complex subject of propagation forecasting can tie their results in with the various types of data transmitted by WWV-WWVH. Those of us who keep schedules across the North Atlantic path look forward to those great N9 conditions.

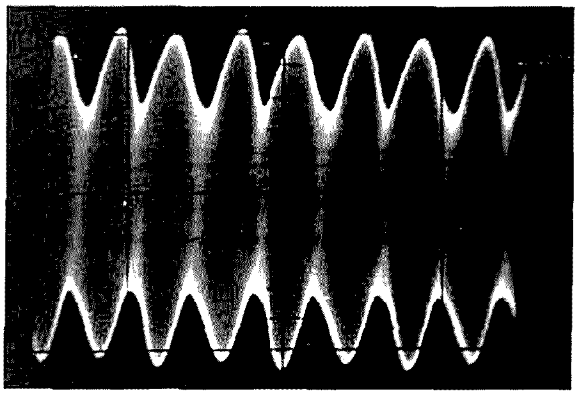
oscilloscopic sideband patterns

Service-type oscilloscopes are available at reasonable cost. Most hams own or have access to such an oscilloscope. Most

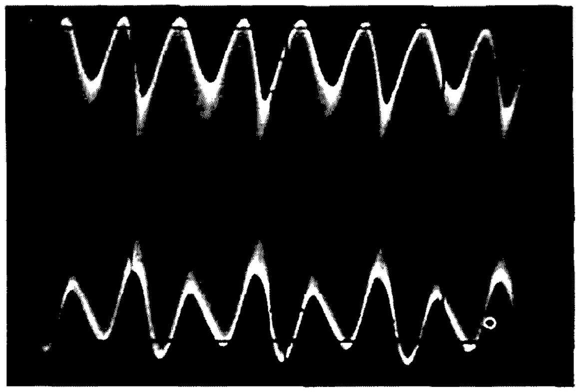


Carrier present.

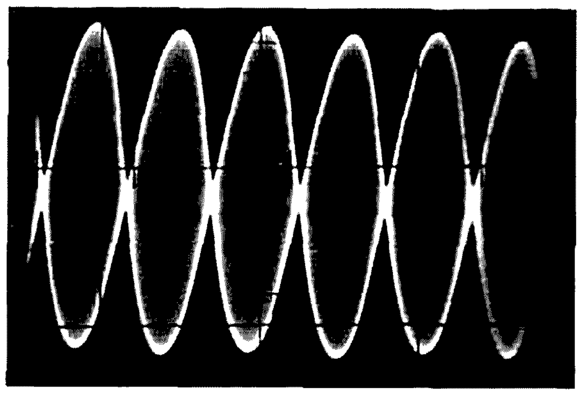
fig. 10. Oscilloscope patterns of a phasing-type sideband generator with single tone modulation.



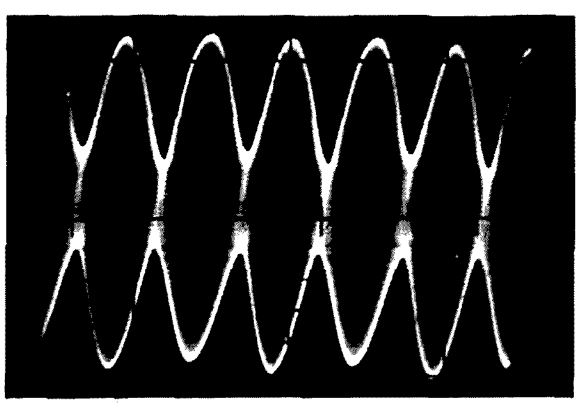
Undesired sideband component.



Carrier and undesired sideband.



Output of balanced modulator shows limited topping.



Output of balanced modulator with carrier feedthrough.

of these instruments have a frequency response in the vertical amplifier system that extends to 4 MHz and higher.

One of the problems of observing and measuring solid-state units is their relatively low voltage and low power level. The high sensitivity of an oscilloscope with a usable vertical amplifier tends to overcome this observation problem. Observations at 160 meters and even 40 and 80 meters are possible.

In the early days of sideband operation much equipment was home-built and many hams were well versed in ssb adjustment, measurement and observation techniques. Solid-state science and QRP activities have brought about a renewed interest.

The set of oscilloscopic displays shown in fig. 10 were photographed from the screen of a service-type oscilloscope. Each of the patterns is labeled and is appropriate for various types of solid-state sideband gear, including a 160-meter phasing-type IC sideband generator to be covered in a future column.

references

1. P. P. Viezbicke, "NBS Frequency and Time Broadcast Services," NBS Bulletin 236, U. S. Department of Commerce.
2. P. C. Lipoma, "$5 Wein-Bridge Oscillator," *Electronic Design*, April 29, 1971, page 65.

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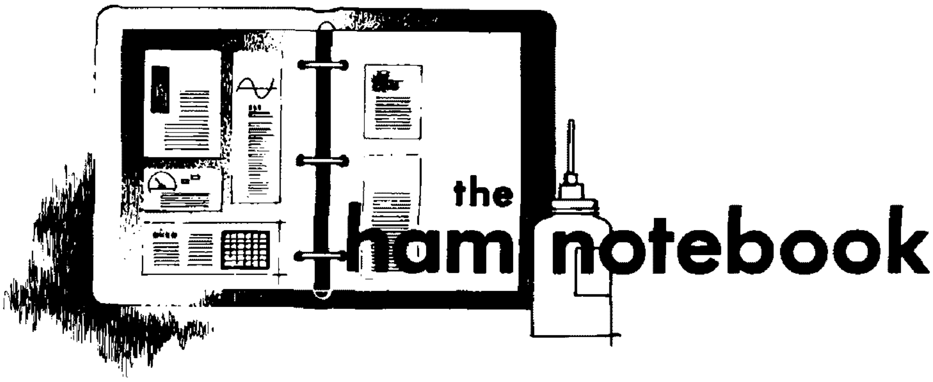


audio filter mod

The circuit in **fig. 1.** is a simple modification of W4NVK's audio filter* that provides variable selectivity with little loss in gain. The 5- to 25-ohm carbon resistor in series with the 88-mH inductor lowers the effective Q of the circuit and broadens frequency response. By adding a rotary switch and several different resistance values you can provide several steps of selectivity. For my purposes one resistor was selected for intermediate bandwidth; a shorting switch removes the resistor from the circuit for sharp selectivity.

Fred Redburn, K6HIU

*E. Dusina, W4NVK, "The Simplest Audio Filter," *ham radio*, October, 1970, page 44.

plug-in toroids

When developing a circuit, you often wish to get it working as soon as possible on one band, with the provision of experimenting on other bands when time permits. A permanent installation is usually out of the question. One simple way to circumvent this problem is to apply a variation of the old plug-in coil technique – use a screw-in toroid.

Two basic parts are the screw-in terminal strip with as many terminals as the tank circuit requires, and a circuit board large enough to accommodate the toroid and its associated capacitors. The leading edge of the board is drilled and filed to accept the terminal strip; the toroid and capacitances are mounted; copper strips are etched to make the connections at the terminal-strip mounting screws. I have used this idea in my direct-conversion receiver and I can easily extend the receiver's coverage with a new screw-in toroid.

It is difficult to approximate the mechanical stability of this setup in touchy oscillator circuits – the thing is as solid as the chassis. With sturdy construction this technique offers better matching in multiband rf finals and drivers. Optimum matching and loading is impossible on different bands by the usual method of tapping down on coils or shorting out sections. The leads associated with the rf tanks in transistor stages must be kept short and as low-loss as possible. You can plug in the optimally tuned circuit for each band without the losses and leads associated with band-

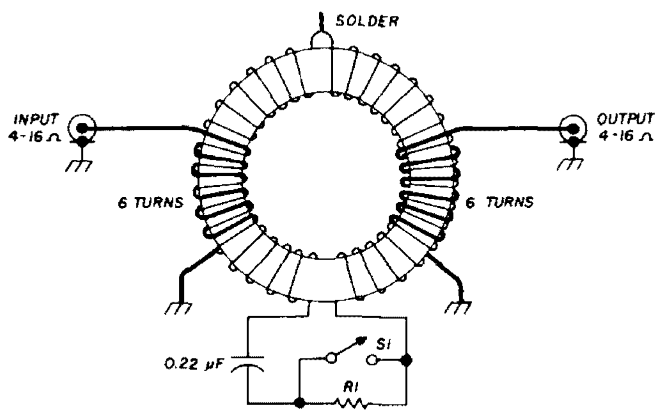


fig. 1. Selectivity of simple audio filter is changed by adding ½-watt resistor, R1, in series with inductor. Value of R1 is 5 to 20 ohms, depending on desired selectivity. Selectivity is broad with S1 open, sharp with S1 closed.

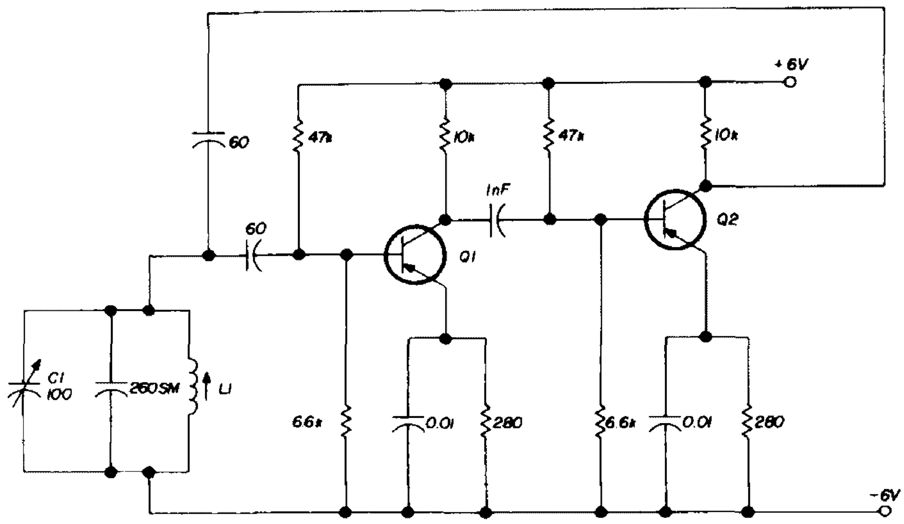


fig. 2. Franklin oscillator using pnp transistors. Use a stable, double-bearing variable capacitor at C1. L1 is 25 turns no. 30, closewound on 3/8-inch iron-core form.

switches and multiband design compromises. On some popular matching systems such as the double-pi section it is easy to lose efficiency by using conventional turns-shorting techniques. With plug-in toroids you get maximum efficiency on each band — this particularly important to the QRP enthusiast.

Adrian Weiss, K8EEG

franklin oscillator

Although known for many years, the Franklin oscillator circuit is seldom found in amateur-built gear. Yet it offers one of the most stable and at the same time most fool-proof circuits available. Nothing about it is critical. It is almost impossible to foul it up to the extent that it won't work! There is no tap on an inductor to juggle for proper feedback. There is no capacitance-dividing network to complicate computation of L/C ratios.

Usually, when you see the Franklin oscillator circuit, it is used with high-impedance active devices, vacuum tubes or field-effect transistors. It will work, though, with bipolar transistors by revising the size of the two capacitors that serve as both feedback devices and isolators. (When used with vacuum tubes or fets, these capacitors are very small, about 2 pF, and therefore serve to isolate the tuned circuit from the possible variations of internal capacitance in the tubes or fets.) With low-impedance active devices, such as bipolar transistors, these capacitors must be increased to around 60 pF to maintain proper feedback. This larger capacitance does not seem to have any deteriorating affect upon frequency stability.

The Franklin circuit, as you'll recognize from the diagram, is closely related to the multivibrator. In fact, that's what it is, with the addition of a tuned circuit to encourage oscillation on a single frequency and with reasonably good waveform. It will oscillate quite readily if you use *hot* rf transistors; don't try to use transistors not designated for rf. If you feed the circuit with regulated dc frequency stability is quite remarkable.

The fourth harmonic of an eighty-meter oscillator, beat against the carrier of a Voice of America propaganda broadcasting station a little above 15 MHz, retained the same audio beat note for hours at a time!

There's little to be said about construction technique, other than the components involved in the Franklin circuit lend themselves to an orderly array on a perforated board; the physical layout is almost exactly like the schematic diagram. Other than one lead, which runs the full length from back to front, there are no cross-overs. A printed circuit

devotee should be able to defeat that cross-over by a long end run!

As with any high-stability low-power oscillator it is best to extract the excitation voltage through an emitter-follower or source-follower stage to avoid disturbing external influences. The buffer stage can be attached to the collector of either transistor. Use as small a coupling capacitor as possible to reduce loading effects.

Carl Drumeller, W5JJ

paralleling power transistors

Theoretically speaking, whenever the power requirements of a particular design cannot be satisfied by a single transistor, several transistors can simply be paralleled to meet the power specifications. In practice, however, this is not always true. Fig. 3 will help explain this.

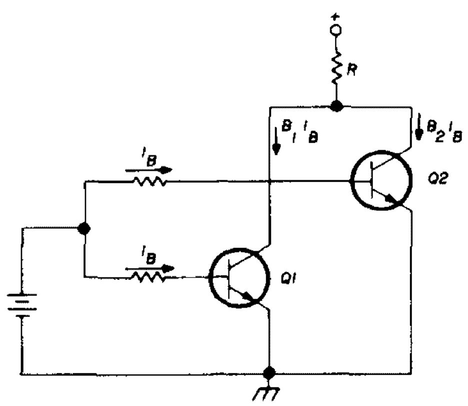


fig. 3. Parallel power transistors Q1 and Q2 share the current through resistor R.

Assume that two transistors, Q1 and Q2, are parallel connected to carry the current through load resistor R. If both devices receive the same base drive I_b, the collector current through each transistor will be determined by its own current gain (B). For many transistor types, the current gain of a given device can differ by as much as a factor of 10 from another device of the same type and still meet the manufacturer's specifications. If, for example, Q1 has a current gain five times as large as the B of Q2, then Q1 will conduct five times as much collector current as Q2 if both receive the same base current. Obviously, a condition of this sort could lead to the eventual destruction of both devices.

In addition to variations in current gains, variations in base-emitter voltages also affect the current conducted by selected devices. As shown by fig. 4, V_{be} partially determines the base current received by a transistor, and this current produces the collector current. Differences in base emitter voltages (V_{be}) between paralleled devices, therefore, will

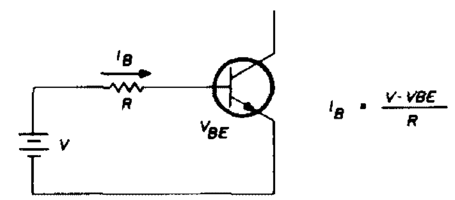


fig. 4. Base current is determined by supply voltage V, emitter-base voltage Vbe and resistance R.

affect the load currents carried by the individual devices.

Identical transistors can be paralleled without an unequal sharing of load currents. Identical devices can be purchased, but they are usually quite expensive in relation to the common garden variety devices. Transistors found on the same integrated circuit chip, incidently, are many times almost perfectly matched and can be easily paralleled.

The adverse effects induced by differences in both current gains and base-emitter voltages in paralleled transistors can be minimized by addition of negative feedback to each individual circuit. This is most easily accomplished by simply placing a resistor in each emitter circuit — the larger the resistor, the smaller the current imbalances in the devices. A point will be reached quickly where further increases in the size of the emitter resistors will not appreciably decrease the current imbalance in randomly selected devices.

Fig. 5 is included as an aid in selecting

the proper value of resistance. According to Smith,[1] this curve is based on statistical data obtained from randomly selected silicon npn transistors. Bear in mind that the curve is approximate and should be used only as a guideline.

As an example, consider that two parallel transistors are to share a 20 ampere load current equally (10 amperes apiece) and that each has an emitter resistance of 0.25 ohms. From fig. 5 (curve A), it can be seen that a current imbalance of 10% is possible. Since 10% of 10 amperes is 1 ampere, one of the transistors could conceivably conduct 11 (10 + 1) amps. Without the resistors, an imbalance of 35% is possible, and one of the devices could see 13.5 (10 + 3.5) amperes.

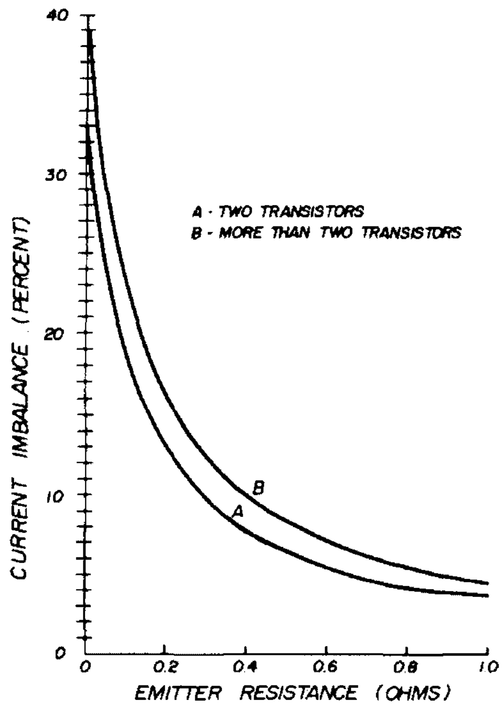


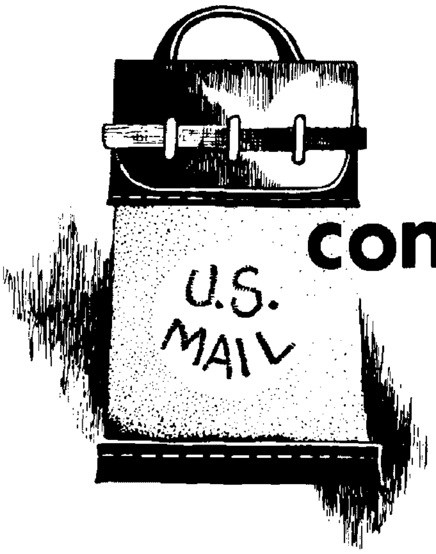
fig. 5. Graph for determining value of emitter resistor when placing power transistors in parallel.

If more than two transistors must be placed in parallel, curve B should be employed.

James E. McAlister, WA5EKA

reference

1. Robert S. Smith, "Selecting Current Sharing Resistors for NPN Silicon Power Transistors," *EDN Magazine*, March 15, 1969.



comments

army loop antenna

Dear HR:

The Army loop antenna mentioned in the September, 1971 issue (page 59) and the method of tuning it seems rather lossy and not very rugged mechanically. This is especially true since a good solid capacitor is already built into the antenna.

When coaxial cable is used for loop antenna construction the capacitance between the inner and outer conductors can be used to advantage both mechanically and electrically. A balanced loop such as the one described could be constructed leaving the inner conductor floating and sealed at the feedpoint. Prune-tune as required from the center equally in both directions. Since an unbalanced coaxial feed does not do too much for a balanced antenna it might be just as well to use one side of the coax loop for one feedpoint and the other side for the other, and do all of the prune-tuning at the base where it is handy.

There is one limitation to this type of loop construction: the inductance and capacitance of the coaxial cable used, i. e. the diameter of the loop and the type of the cable. For a quick check I took a 38-foot section of ½-inch semiflexible line (25 pF per foot) which ran in a single-turn loop and grid-dipped it at about 10.5 MHz. A 7-MHz loop antenna with the inner and outer conductors of the same length are a practical size and could be used on 75 meters with a plug-in series capacitor.

Since short loops have quite low radiation resistance, the resistance of all conductors and connections must be kept to a minimum. Coaxial-cable construction should minimize (or eliminate) the need for additional tuning units that are lossy, bulky and do not stand the weather very well. Vinyl-covered coaxial cable is well protected from corrosion and the weather, and the ends can be sealed and painted with corona reducing material; rf power-handling depends upon the size of the cable. There are a number of easy-to-form semiflexible cables from which to choose.

This type of coaxial cable also makes a good solid weatherproof gamma rod capacitor for feeding parasitic beams because it eliminates the corroding slide rod and variable capacitor. Prune-tuning the coaxial-cable gamma matching system may take a little longer but is well worth the effort when it comes to durability. I used such a gamma match on a recently built full-size loop antenna (one wavelength long) to good advantage; the winter snow and icing are rough down here!

Wayne W. Cooper, K4ZZV
Miami Shores, Florida

1296 preamplifier

Dear HR:
In WA2VTR's "Low-Noise 1296-MHz Preamplifier" article in the June, 1971 issue of *ham radio* the author referred to Hewlett-Packard 5082-2800 hot-carrier diodes as mixers at these frequencies. This device is a good high-level mixer, but the HP 5082-2835 is a much better low-noise uhf mixer.* They are low cost, too.

Leonard J. Petraitis, W9IKW
Regional Sales Manager
Hewlett Packard
Skokie, Illinois 60076

infrared communications

Dear HR:

In "Gallium Arsenide LED Experiments" in the June, 1970 issue W4KAE describes some equipment for using infrared light-emitting diodes and photodetectors for communications. I've experimented since about 1969 with LEDs but the big problem has been building lens systems which are rigid enough to stand up under continuous use, yet allow easy adjustment. This is the hardest part if you don't have a fully equipped machine shop. I recently came across a piece of equipment which, though designed for something completely different, can easily be converted for communications. A photoelectric cell system designed for burglar alarms, and made by the Alarm Device Manufacturing Company, 165 Eileen Way, Syosset, Long Island, New York 11791.

One version of this series of burglar alarm equipment is the model 1312 *Small Laser Photoelectric System* which actually uses a LED pulsed at several kHz, and a silicon photodiode detector. It comes in two units, one with the LED and the other with the detector. In

*In addition, the HP 5082-2835 has a typical capacitance of 0.75 pF as compared to 2.0 pF for the 5082-2800. Therefore, it is less apt to operate as a varactor.editor

burglar alarms it has a guaranteed range of 800 feet (assuming no fog or rain), but most units can be used for quite a bit more. Both units are complete with mounting hardware, optics, an easily adjustable swivel arrangement and case, and are designed for installation by inexperienced installers. To convert it to communications it is only necessary to modulate the pulsing circuit to produce pulse frequency modulation and detect the fm in the output.

For serious experimenters the price of $125 may make the system of some interest, but it is hard to get, since the manufacturer sells only to burglar alarm installation companies. If your local alarm installer refuses to sell you one, or if he wants too much of a markup, I can get a limited number at the above price, including shipping.

Peter A. Stark, K2OAW
Mount Kisco, New York

RTTY afc

Dear HR:

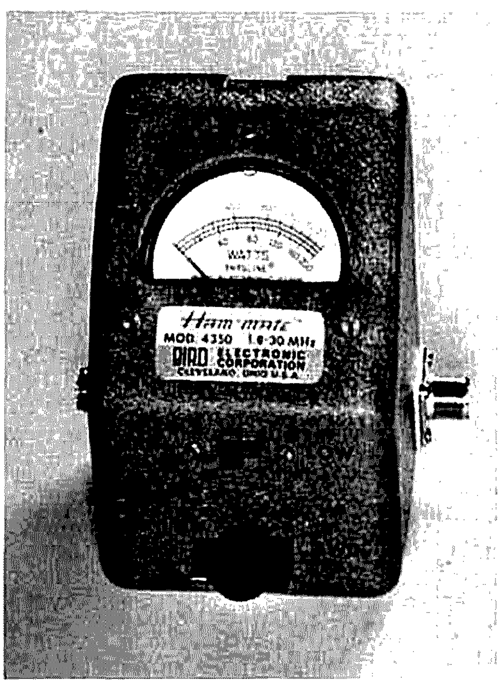
While reading the article, "Automatic Frequency Control for RTTY" in the September, 1971 issue, I came across one point that should be clarified for the benefit of your readers. In the article W5NPD states that in most RTTY signals 2975 Hz is used as the mark tone and that this tone is present for a larger percentage of the time. It should be noted that the standard mark tone has been 2125 Hz for years; most, if not all, AFSK systems use 2125 Hz for mark. Many builders have omitted a normal/reverse switch and straight wired the unit in the normal 2125-Hz position.

To use 2975 Hz as the mark tone the operator should listen in the upper sideband (bfo low) position with the normal/reverse switch on the converter in the reverse position. For those who use their converter in the normal (bfo high, lower sideband) position, it is feasible to build the afc using 2125 Hz as the standard for the discriminator tuned circuits.

Tony King, WA4UPE
College Park, Georgia

new products

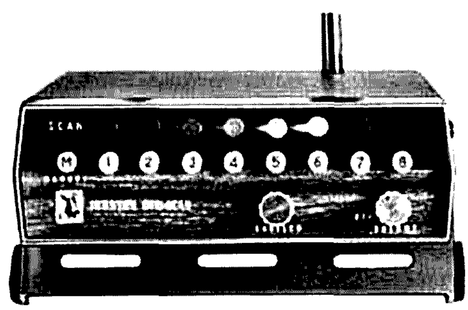
ham wattmeter



Designed especially for the amateur radio market, three new rf wattmeters are now available from Bird Electronic Corporation. Two models will cover the frequency range from 1.8 through 30 MHz and the third covers from 50 to 150 MHz. The Model 4350 measures forward and reflected power in two ranges: 200 w and 2000 w, while the Model 4351 has ranges of 200 w and 1000 w. The Model 4352 has ranges of 40 w and 400 w covering the two vhf bands of six meters and two meters.

The new line of wattmeters are designated Ham-Mate,* and use the well known Thruline construction, made famous in the industrial field by the Bird Model 43. The new wattmeters emphasize dependable rf power measurement. Special attention is given to the directivity of the Ham-Mate, which is the ability to differentiate between rf power flowing in opposite directions in a transmission line. The new Ham-Mate has a minimum of 20-dB directivity which assures meaningful reflected power and vswr measurement. All three models sell for $79. More information is available by using *check-off* on page 94, or by writing to Bird Electronic Corporation, 30303 Aurora Road, Cleveland, Ohio 44139.

scanning receiver



A new fm monitor receiver which covers both low- and high-band vhf channels has been introduced by the E. F. Johnson Company. Called the Duo-Scan* monitor receiver, it features dual-conversion circuitry and a double ceramic filter. The two ceramic filters result in sharp selectivity characteristics and an adjacent channel rejection of 60 dB. Another feature of the new Duo-Scan is an integrated-circuit limiting system. This IC limiting produces a symmetrically *hard*

*Ham-Mate and Thruline are registered trademarks of the Bird Electronic Corporation.

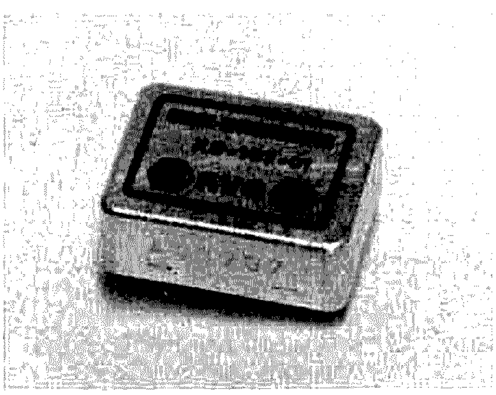
More Details? CHECK—OFF Page 94

pattern that effectively eliminates noise. The receiver's all solid-state circuitry has a sensitivity of 0.4 μV for 12 dB SINAD.

The Duo-Scan receiver has eight channels which can be any low- and high-band combination. If just one low-band frequency is to be monitored, for instance, the remaining seven channels can all be used for high-band monitoring. This allows you to monitor all the local 2-meter repeaters and the activity on 52-525 MHz. Programming channels for high- or low-band is done with simple jumper plugs, requiring no rewiring or tuneup adjustments. Each channel has a lock-out pushbutton to permit bypassing of that channel if desired. The monitor can also be used in the manual mode with the pushbuttons used to lock in channels.

A built-in power supply allows both 12-Vdc mobile operation and 117-Vac base-station operation. Other features include noise-operated squelch, channel selector indicator lights and external speaker jack. Suggested price for the receiver is $169.95. Complete details are available from local E. F. Johnson dealers, by writing directly to the E. F. Johnson Company, Waseca, Minnesota 56093, or by using *check-off* on page 94.

crystal discriminators



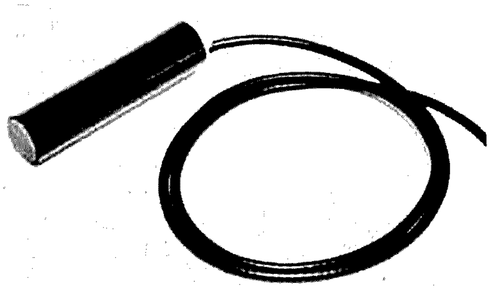
Spectrum International has added five new crystal discriminators to their line of crystal products manufactured by KVG in West Germany. The new models, XD 9-01, XD 9-02 and XD 9-03 are designed for RTTY, afc and lab measurements with a 9-MHz center frequency, the model XD 107-01 for narrow-band fm

and the model XD 107-02 for wide-band fm. Both fm versions have a center frequency of 10.7 MHz.

Crystal discriminators make simple and effective demodulators for fm and RTTY receivers. The receiver i-f output is fed into the discriminator input, and audio output results without the tedious alignment and power requirements of ratio detectors and LC discriminators. Input and output transformers are included in the hermetically sealed crystal discriminator package. Compatible with tube or solid-state equipment, these miniaturized discriminators are designed for replacement in high-density conventional equipment along with new miniaturized designs.

The line of five filters include models with deviation between 5 and 50 kHz. All models have standardized input terminations of 500 ohms and outputs of 100 kilohms. Models are priced at $14.95 and $16.95. Complete technical specifications are available from Spectrum International, Box 87, Topsfield, Massachusetts 01983 or use *check-off* on page 94.

sniff-it rf detector



The Dycomm Sniff-It turns any vtvm or vom into a versatile rf detector. The model DS-1 Sniff-It indicates the presence of rf from the lowest intermediate frequencies up to uhf at power levels ranging from one milliwatt to 250 watts. A typical reading with the Sniff-It connected to a Simpson 260 vom on the 50-μA range would be 5 to 10 μA when placed near a 1-mW 450-MHz oscillator.

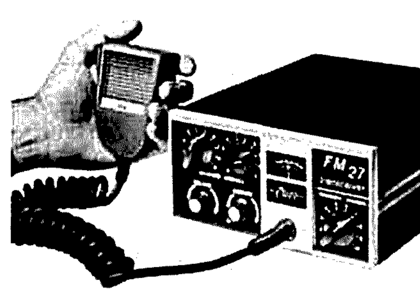
Uses for the Sniff-It include troubleshooting receivers (by indicating the presence of signals near the rf and i-f stages and output from the local oscillator), aligning converters, peaking crystal oscillators and QRP projects and balancing push-pull amplifier stages (by placing the Sniff-It next to each stage and adjusting until equal readings are obtained). When tvi proofing the station, the Sniff-It detects rf leaks around the transmitter cabinets, feedthrough to cables, coax shield breaks leaking rf, local oscillators pumping out too much rf and unneutralized rf stages.

The Sniff-It can also be used in adjusting antennas. You can run the Sniff-It along a ¼-wave whip and check for nulls. If the null is at the end of the antenna you know you have the right length. If the null comes before the tip of the antenna you know that the antenna is too long. Similarly, antenna feedlines and 5/8-wave-length verticals can be adjusted by checking for nulls and comparing your findings to the theoretical figures found in any good antenna handbook.

By choosing the current range on the vom, the Sniff-It detects a wide range of power levels. A vtvm can be used to show relative voltage drops across a load resistor.

The Sniff-It is available for $5 from Dycomm, 948 Avenue E, Riviera Beach, Florida 33404. More information is available by using *check-off* on page 94.

fm transceiver with frequency synthesized receiver



Clegg's nwe FM-27 is a completely solid-state mobile transceiver using a *Crystiplexer* tuner. The *Crystiplexer* is a new synthesizing system that allows any channel in the range of 146-147 MHz to

More Details? CHECK—OFF Page 94

be monitored with crystal precision — but without the need for additional crystals. To monitor any specific frequency within this band the operator merely sets the two receiver controls to numbers corresponding to the desired frequency. To monitor 146.94, for example, the operator sets the first control to 9, and the second to 4.

Receiver selectivity is rated at 70-dB adjacent-channel attenuation. Sensitivity is rated at better than 0.35 μV for 20-dB quieting.

The transmitter portion of the new transceiver is 10-channel solid-state unit with 20-25 watts rf output. The entire transceiver weighs less than three pounds and is packaged in a rugged anti-theft case with a special locking-clamp mounting. Two crystal-controlled transmit channels and a push-to-talk microphone are included at the sales price of $449.95.

Further information is available through *check-off* on page 94 or by writing to Clegg Division, International Signal and Control Corporation, Box 388, R. D. 3, Lititz, Pennsylvania 17543.

tube and transistor substitution guide

Tab books has brought out the updated 3rd edition of their handy servicing substitution guide. The 256-page "1972 Popular Tube/Transistor Substitution Guide" is designed to fit in a tube caddy and lists 99% of the tubes and transistors which normally need replacing in home-entertainment equipment. Moreover, only readily-available and comparably-priced substitutes are listed — no need to search through lists of tubes and transistors rarely used.

This brand-new volume contains 8 sections — four devoted to tubes and four to transistors. Section 1 provides a cross-reference of popular American receiving tubes, listing substitutes which have similar or superior characteristics and which require no mechanical changes or circuit modifications. For convenience, the *best* substitute is listed separately from others that may be used. Section 2 lists substitutes for popular tube types found in commercial and industrial equipment. Again, the most appropriate substitute is identified. Section 3 provides a cross-reference of popular foreign/American tube types, and Section 4 illustrates base diagrams for all tubes.

Section 5 contains a complete listing of popular American transistors and the most readily-available, popularly-priced substitutes. Section 6 lists American substitutes for the most often encountered foreign transistors. Section 7 lists general-purpose replacements, offered by leading manufacturers. Section 8 illustrates base diagrams keyed to the original type listings.

The vinyl-covered version of the book costs $4.95 and the paperback is $2.95 from Tab Books, Blue Ridge Summit, Pennsylvania 17214. More information is available by using *check-off* on page 94.

audio ics

Radio amateurs and experimenters may benefit from two new audio integrated circuits introduced by Motorola. The two new ICs, while designed for such consumer product applications as televisions, radios and portable phonographs, should also work well in many amateur receiver and experimental designs.

The MFC9020 is rated at two watts output and is housed in a plastic package with two heat dissipating tabs. The MFC6070 is the one-watt version and is supplied in a smaller case. Printed-circuit board layouts which can be used interchangeably with either device, are shown on the respective data sheets. The boards are designed for speaker return to either ground or to the positive supply. Input impedance is in the order of a megohm, and only 200 mV input is required for full output. Total harmonic distortion averages about 1% at rated output.

For further information, please contact the Technical Information Center, Motorola Inc., Semiconductor Products Division, Box 20924, Phoenix, Arizona 85036 or use *check-off* on page 94.

More Details? CHECK—OFF Page 94

reference data for radio engineers

Whether you are in amateur radio, commercial radio or electronics, the new fifth edition of "Reference Data for Radio Engineers" is an invaluable tool. This new compilation of approximately 1150 pages and 1350 illustrations plus a comprehensive index and cross-index was made by an extremely qualified group of practicing engineers, professors, and industry and government experts under the direction of the staff of the International Telephone and Telegraph Corporation.

The book is skillfully written, greatly enlarged and meticulously revised and edited. Approximately 50% of the text is new material, including seven subject areas not covered by the fourth edition. In addition to the basic phases of electronics, there is new material on microminiature electronics, space communications, navigation aids, quantum electronics and many other topics.

In 1942, the British subsidiary of ITT, Standard Telephones and Cables Limited, saw the need for a complete, reliable reference source for the radio and electronics engineer. Thus, the first edition of this reference was developed as a 60 page brochure. Because of the enthusiastic reception of information compiled under one cover, the brochure has grown into a book-length volume.

Its usefulness, however, is not restricted to the practicing radio and electronics engineer for whom it was originally prepared. *Reference Data for Radio Engineers* has been accepted for classroom use in over 200 leading colleges and universities in the United States.

The text is generously supplemented with literally hundreds of charts, nomographs, diagrams, tables, circuit information, illustrations and an inexhaustible index and cross-index. 1150 pages, hard bound. Published by Howard W. Sams & Company. $20.00 postpaid from Comtec Books, Post Office Box 592, Amherst, New Hampshire 03031.

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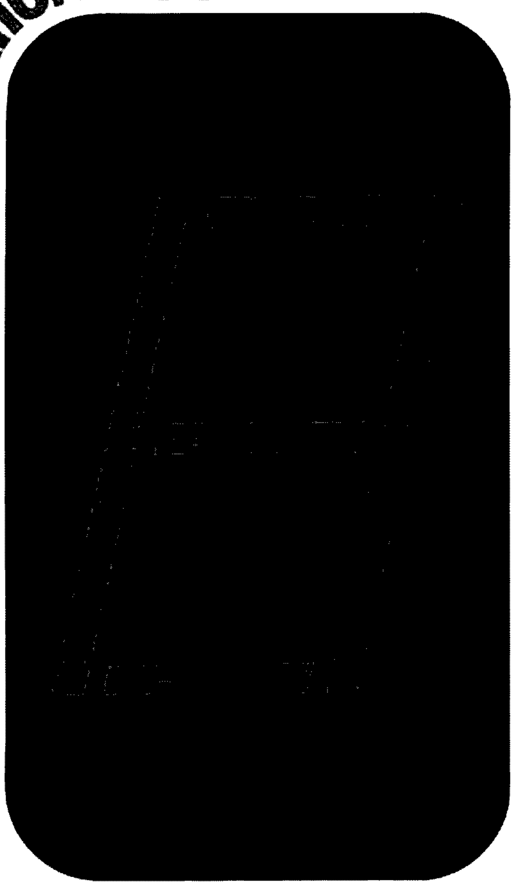
FEBRUARY 1972

DIGITAL READOUT STATION ACCESSORY



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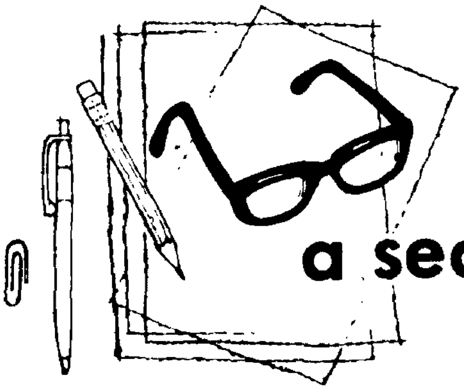

a second look

by jim fisk

As more and more amateurs switch to factory-made gear, and as industry uses more ICs and disposable plug-in modules, the life of the dyed-in-the-wool homebrewer gets tougher and tougher. If you've recently tried any of the construction articles in *ham radio*, you are already well acquainted with the hassle involved in obtaining a few needed components.

At one time you could drop in at your local corner radio store with a list of parts and the man behind the counter would fill your order. But that was when the vacuum tubes, resistors and capacitors in your ham gear were the same as those in the family radio. It's not the same anymore – now the transistors and ICs in radios and television sets are apt to be designed specifically for that purpose and have operating characteristics that are of little use elsewhere. There are exceptions, but they are few and far between.

Another problem that faces the serious home builder is the generous variety of transistors and ICs available from different manufacturers. Although some types of devices are made by more than one company, in most cases the semiconductor manufacturers crank out devices that are completely different from those of their competitors. And to add insult to injury, the same device may carry a dozen different part numbers: a 2N number, a replacement number plus special numbers for units sold in large quantities to equipment manufacturers.

There is only one way to combat this lunacy: arm yourself with a good semiconductor cross-reference guide and a wide selection of electronic parts catalogs.

Tops on the list of replacement guides is Howard Sams "Transistor Substitution Handbook." This handy little paperback, which is updated every year, covers practically every transistor ever made, from 2N34 to 2N6000, with recommended substitutes. It also covers devices from Japan and Europe, as well as replacement types manufactured by Delco, General Electric, International Rectifier, Motorola, RCA, Semitronics, Sylvania and Workman. Most of these manufacturers also publish replacement guides, available for the asking from their authorized distributors.

If you live in a large metropolitan area, chances are that there is an industrial electronics supply house that can fill your parts need. Many of these firms don't advertise, because they are not particularly interested in small quantity sales, but if you show up at their office, they will sell you the parts. If you want to find them, pick up your telephone directory and check the Yellow Pages; look under "Electronic Equipment and Supplies." If you live out in the sticks, the problem is more difficult, unless you can get into the city. If you can't, you must purchase your components through the mail. Allied Radio is the best bet in this case and you can get a catalog from any Allied/Radio Shack store. Be sure you get the Industrial catalog though, the more common entertainment catalog is devoted primarily to hi-fi, CB and simple experimenter's stuff.

Jim Fisk, W1DTY
editor

digital readout

station accessory

This versatile station accessory may be used as a digital vfo readout, digital clock, identification timer, electronic keyer or frequency meter

The most attractive magazine articles may be those that provide instructive or interesting reading to a large percentage of readers, including those readers who do not plan to build or use the device described. It is hoped that this article will be of that type.

E. H. Conklin, K6KA, Box 1, La Canada, California 91011

Possibly each of us should plan to build one electronic gadget every year, just to avoid becoming entirely an "applicance operator." This is not easy to do under present conditions other than for small items of accessory or test equipment. Of all the gadgets other than kits, digital nonlinear integrated circuits offer perhaps the most pleasant areas in which to experiment, design and build an accessory for your station. There is widespread interest in such devices, as will be seen from the many technical papers which have been published during recent years.

Instead of building a series of devices to accomplish different things, I have given consideration to the idea that one cabinet, chassis, power supply and readout can be used for an almost endless number of jobs. This saves construction, space and expense. By using some form of plugboard for the circuitry, new active devices can be added, or replaced even temporarily, such as when you are asked to time a swimming meet or auto race to a thousandth of a second.

I plan to cover the basic unit at this time, and the detailed accessory devices in coming months. Some of the accessories will not be the subject of an article because of the number of similar units already in print. These can be built on suitable plugboards and used in the accessory unit.

In the meantime, general planning and

background reading well may be in order to acquaint you with the subject. This will make the subsequent plug-in devices very easy to understand. In fact, they may not even require a circuit diagram, except for block diagrams and switching plans for the station maintenance files.

Some of the items that can be worked into the accessory are: high-stability crystal calibrator, harmonic generator, frequency counter, digital clock, identification timer, receiver digital dial, transmitter digital dial, general-purpose receiver dial, electronic keyer, event timer or stopwatch, auto speed trap, photo timer and *wife-reminder*. The *wife-reminder* refers to anniversaries, birthdays and the like. These, and license renewal date, can be added, with suitable alarms for the purpose!

In preparing such an article, it is desirable to simplify parts procurement in order to eliminate a month or two of planned purchasing, with the inevitable back-ordering and interdependence of one item on what is obtained for some other requirement. Also, chassis and cabinet dimensions given are for the very few that matter. And finally, difficult machining problems and other sources of frustration and discouragement should be avoided.

nixie read-out displays

Chassis construction depends to a considerable extent upon the type of read-out displays that you select. Tube-types such as Nixie (a Burroughs trade name) and Numatron (RCA) may require a specific chassis height, or shelf, or mounting on the forward edge of vertical epoxy-glass circuit boards. It is feasible to use a chassis only in the center to support the read-out tubes and the power supply with all other parts except panel switches mounted on plugboards (such as Vector *plugbords*).

The widely known Nixie gas-discharge display device requires a few milliamps at about 180 volts. It also requires a high-voltage BCD-to-decimal decoder/driver IC. As in other uses of decoders and drivers, care must be taken to obtain those with *active high* or *active low* operation as required by the particular type of read-out. A current-limiting resistor is used as with neon tubes. Sometimes a grounded wire-mesh screen is placed over the panel window to prevent inside radiation from leaking out, or the reverse.

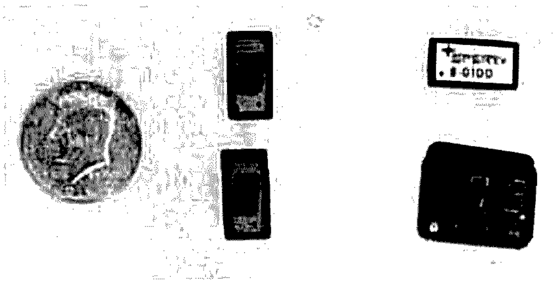

fig. 1. Two Minitron readouts are shown on the left. A three-digit Sperry cold-cathode display, resting on its special socket, and one of the decoder/driver ICs, are on the right.

These read-outs will be found in large mail-order industrial catalogs. However, note that this is a fast-moving field, and spare tubes should be on hand to prevent running short if the type you use is not available in the future.

Burroughs has extended the cold-cathode gas discharge design into *Panaplex,* a single tube with eight or more 7- or 9-segment digits, including driver circuitry. A strobed or multiplexed time-sharing method is used to minimize decoder/driver cost. The hoped-for cost of $1 per digit is a few years away.

In the meantime, Sperry Information Displays* is producing some very pretty replacements for Nixie tubes. These are 1½-, 2- or 3-digit flat glass plates, a socket and active-low 7-segment decoder-driver units, shown on the right in fig. 1. Sperry offers demonstration kits including read-out, socket and decoder/driver, and may furnish the 2- or 3-digit ones on request at $14.95 for the lot, thus being one of the low-cost devices. They do

*Sperry Information Displays Division, Box 3579, Scottsdale, Arizona.

require the low-current 180-volt supply, though.

General Electric offers the Y-1938 7-segment low-voltage tubes with filament. They have an unusual blue-green character color and require 1.6 volts on the filament and 25 volts dc on the anode, but can be driven by suitable IC decoder/drivers. General Instrument and Hughes make up/down counters, buffer storage, and decoder/drivers in a single large 24-pin dual in-line package (DIP) IC that will operate the GE tube. These combined counter ICs have merits but are not desirable for digital clock purposes.

incandescent displays

The RCA Numitron is a 7-segment incandescent tube, taking up to 5 volts, but at 24 milliamperes per segment. This requires one of the high-current decoder/drivers.

There are other simple incandescent lamp displays, some requiring higher current than is provided by a 7-segment decoder/driver IC. One made in Japan, however, built like a 16-pin DIP IC, is the Minitron, shown on the left in **fig. 1**. Its 5-volt and low-current requirements are well within normal 7-segment decoder/driver output, like the inexpensive active-low Monsanto MSD 047 and the Texas Instruments 7447, all available from Circuit Specialists.* These have provision for decimal, lamp test and read-out blanking input and output. Currently the Minitron and MSD 047 decoder/driver combination appears to be the lowest-priced read-out system available.

light-emitting diodes

The Monsanto MAN3A light-emitting diode read-outs are built somewhat like a ten-pin IC flat-pack, with pins spaced 0.034 inches. The 7-segment display is only 0.115 inches high, thus about the smallest in the business. At around $6.50 a digit this is the lowest cost of the bright red-colored light-emitting diode displays. The Heath 14-pin DIP LED display, used in their 80-MHz counter, sells for $8.50.

Light-emitting diodes are attractive and will be very popular when the small-quantity cost comes down a bit. They require a dropping resistor in each segment lead and each decimal-point lead. LEDs are compatible with systems that include the decoder/driver, latch, and possibly the decade counter, all in the same IC.

other displays

Sigmatron* has a light-emitting film multidigit display; in large quantities, this runs only $1.50 a digit, but unit prices are much higher. These units require an audio-frequency multivibrator driver which can be time-shared by all of the digits with a multiplexing device.

Optel† offers liquid crystal displays. These require a small 20-volt supply. With the decoder/driver and Hughes HCTR 0107 display driver the small-quantity cost is much higher than many other read-outs.

color filters

Usually a color filter is placed over a numeric display to increase the contrast between the read-out's light and the background. Most displays are red, but a few are blue-green. The incandescent ones are close to an orange-white color, depending upon the voltage at which they are operated.

Litronix mentions that the Rohm & Hass no. 2423 acrylic plastic has 71% light transmission, and that the Polaroid HRCP-7 has only 40% light transmission but offers a nonglare surface. In a test, Circuit Specialists did not find any advantage in the Polaroid material compared with sticking red Scotch tape on individual Minitrons and trimming the edges even.

Plain half-sheets of thin flexible plastic cost about $1.50 at art supply stores; similar colors are available at school supply stores as "report covers" at 19¢

*Circuit Specialists, Box 3047, Scottsdale, Arizona 85257.

*Sigmatron, 849 Ward Drive, Santa Barbara, California 93105.

†Optel, Box 2215, Princeton, New Jersey.

each, which is quite a bargain. The red looks good. For incandescent displays, the rose (not quite amber or orange) probably passes more visible light, even when two thicknesses are used.

It is desirable to minimize light spread by having the plastic window close to the read-out. The chassis described in fig. 6 has the plugboard holding the displays directly behind the panel window with the Minitrons actually touching the plastic. The plastic can be glued to the rear of the panel or it can be secured with flat-head 2-56 machine screws with a washer and nut on the rear side of the panel.

For trying different colors, a 3 x 5-inch card holder can be purchased at a builders' hardware store. This can be held in place on the rear of the panel with flat-head machine screws. It is practical to cut out a section on the 3-inch dimension, mounting the old top and the remaining side lengths from the upper machine screws. This has been done on the panel shown in figs. 2 and 6. Inasmuch as the Minitron read-outs are thicker than the other ICs there is no problem in clearances from the panel. If fewer than eight read-out digits are planned, such as by omitting the megahertz, smaller escutcheons can be used without modification.

number of read-outs

The unit pictured in fig. 2 has eight Minitron read-outs in one row, shared for all purposes, so that 99 MHz can be displayed right down to the last cycle. In some applications it is feasible not to bother with counters, latches and decoder/drivers for the MHz digits; if these are available they can be activated by a bandswitch to connect 5-volts dc through steering diodes to the correct read-out segments.

Some read-outs can be omitted, such as the MHz ones, or the units or tens of Hertz. It is quite useful to have only three kilohertz digits and one for tenths-of-kHz, thus requiring only four read-out tubes.

Another option is to use two rows of read-outs, with frequency (counter) on one row, and time of day on the second row — if sufficient plugboard contacts are available. In this way, by not sharing these displays, there will be the added cost of the extra read-outs and their decoder/drivers for the extra four units or more used to display time.

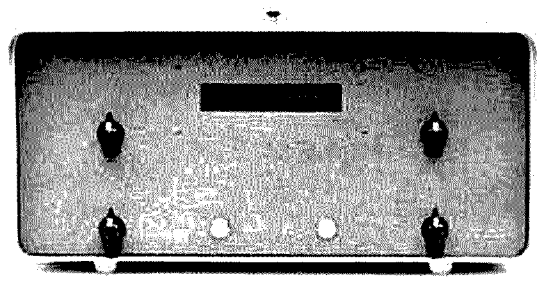

fig. 2. Front panel and cabinet of the digital accessory prior to the application of panel lettering.

There will be some saving on switching, and also about $5 for the four quadruple 2-input AND gates used as data selectors on the displays alternately used for time and for counting. Texas Instruments promises production of their SN74157N quadruple 2-line-to-1-line data selectors to do this same job of switching between two types of input data to the read-outs, but prices are not yet available.

cabinet and chassis

With IC-type read-out displays a miniature unit is possible, but it limits flexibility in usage and may complicate experimentation, test and maintenance. Readers with other Heath equipment can select from a number of cabinets used in many Heath units.

As a first attempt at packaging, I decided to use the gray CO-1 cabinet made by LMB shown in fig. 2. It measures 14½-inches wide, 6½-inches high and 13½-inches deep including the shadow front. It closely matches Collins S-Line cabinets and the original Henry 2K.

Through separate order, the alternate 1½-inch-high chassis – not the more popular 2-inch-high chassis – was ordered. These are obtainable through supply stores. Unlike the Collins cabinets, it does not come with a rectangular hole for access to the rear apron of the chassis. Such a large rectangular hole can be cut out, or small guide holes can be drilled from the rear of the cabinet into the rear of the chassis and later enlarged to pass phono plugs through the cabinet into phono jacks mounted on the chassis.

This operation is facilitated by shimming the rear of the chassis 1/16-inch with washers held in the corners with masking tape or by a 4-40 machine screw permanently threaded into each of the two rear chassis angle brackets. Scrap pieces of Vector T-strut can be used to hold the chassis by placing these pieces in the guide rail at the sides of the chassis. By blocking up the chassis from the front and holding it from popping out, guide holes for the entire set of phono jacks can be drilled.

Subsequently, sets of black and white letters* can be applied to both the gray cabinet and the aluminum chassis to identify the phono plugs. The Letraset is smaller than that offered by electronics mail-order houses; it and other alternatives are available from art supply stores. Crystal-clear Krylon spray coating over the transferred letters will make them permanent.

There is no provision for a chassis fuse nor for a complete on-off switch inasmuch as the oven and clock components will remain turned on except during trips abroad. The unit is fused in the power plug (Allied Radio Shack no. 270B1249).

The large hole required for a power-cord strain relief is beyond many home drilling facilities; a grommet with suitable tight-fitting hole will do a reasonable job. Stretch the grommet onto the power cord, then drill out the chassis hole that will accept the grommet but retain the power cord.

The cabinet's cover has no provision for a handle or finger hole. However, a 4-40 machine screw with fibre washers and spacer under the head, will pass through a ventilation hole, to be used as a cover handle.

If the unit will be near Collins equipment, obtain two or four tapered feet (Collins 543-8101-002). Four are used

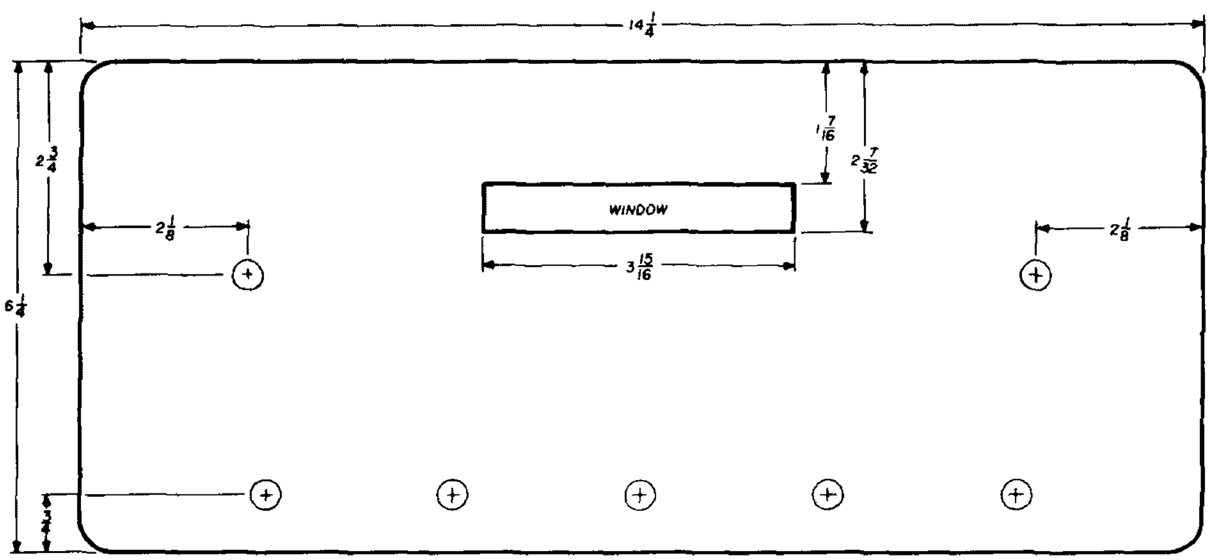

Fig. 3. Front panel drilling instructions.

*Letraset 10-point Helvetica light (1573), catalog number 48-10-CLN, upper and lower case, available at art supply stores.

when stacking above or below S-Line units, maintaining ventilation space. Rubber feet are Collins 200-5010-000.*

the panel is the right distance back at its bottom, then drill new screw holes up through the bottom. A sheet-metal screw

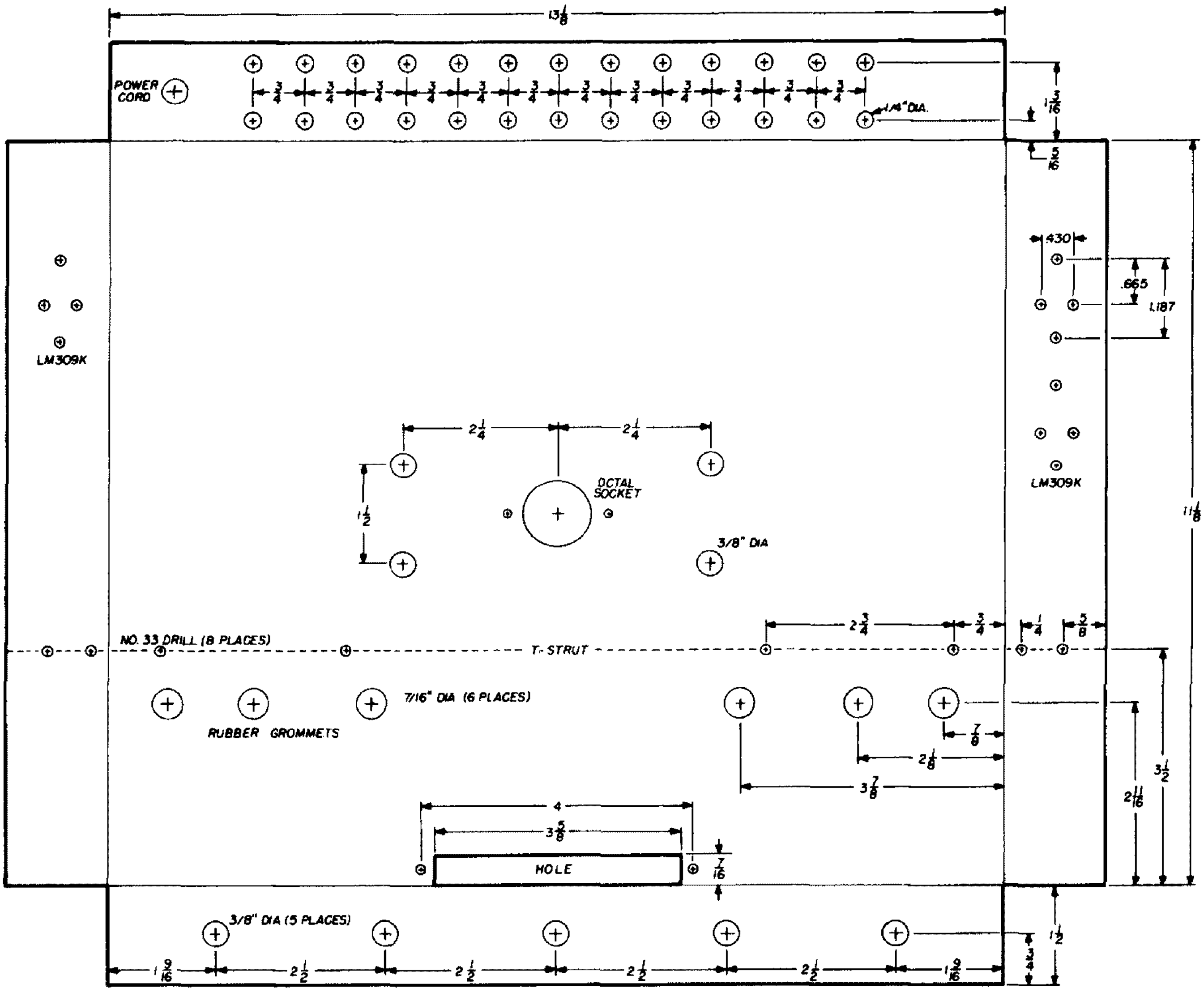

fig. 4. Chassis drilling instructions; the power-supply holes are omitted because they will not apply for different parts.

panel

The front panel angle comes secured under the cabinet cover by two sheet-metal screws which also hold down the hinged top. This is inconvenient for adjusting the crystal oscillator and setting the clock. The answer is to clamp a piece of wood to the lower cabinet lip so that

*Collins parts may be ordered from Customer Service Representative, Collins Radio Company, Cedar Rapids, Iowa 52406.

should be put into the first hole before the second is drilled. In this way, the chassis and the front panel are held firmly, even when the cabinet lid is opened.

The lower set of controls will require holes ¾-inch up from the panel bottom, through the chassis, as shown in figs. 3 and 4. Old controls or their bushings then can be inserted to hold the chassis and panel in place as additional holes are drilled.

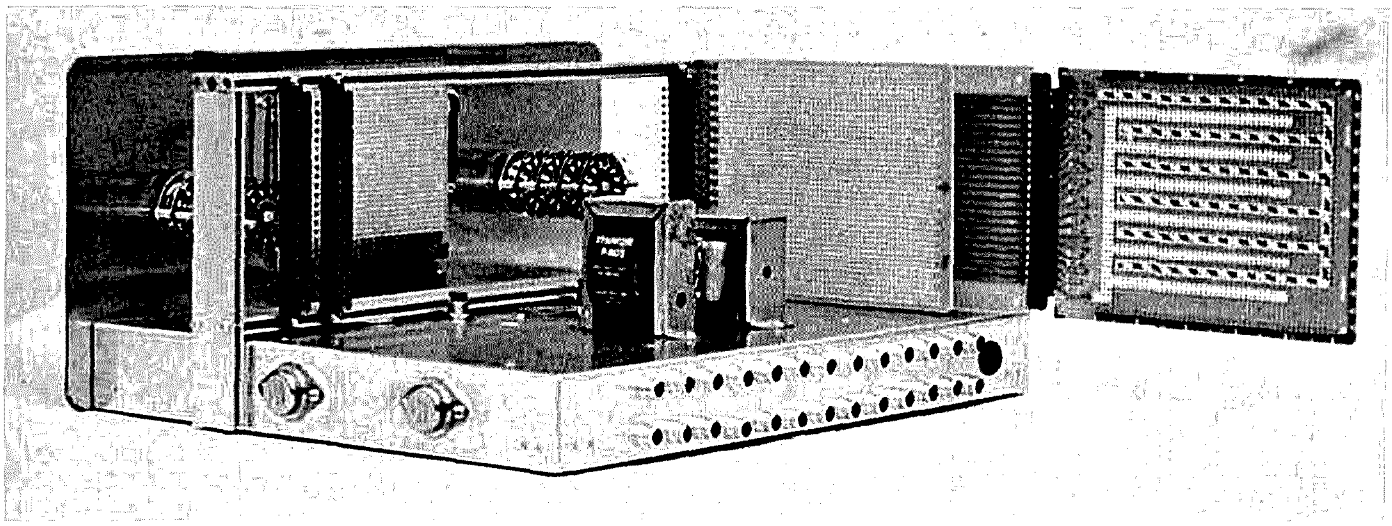

fig. 6. Rear view of digital accessory showing the mounting of Vectorboard to hold Minitron readouts and T-struts for other circuit boards prior to above-circuit chassis wiring.

It can be seen in fig. 5 that the small space under the chassis permits a switch only as large as the Centralab PA-1000 and PA-2000 subminiature series. Even then the switch must be wired when dismounted, bending contacts in, and applying plastic tape to the chassis and cabinet to prevent the contacts from shorting.

These switches are attractive because spare parts and wafers are available for possible future expansion. Smaller switches may require smaller panel holes and 1/8-inch shafts. The small Daka-Ware knobs that were supplied with the switches are helpful when positioned so close to the bottom of the cabinet.

Some controls can interfere with the read-out plugboard receptacle contacts. In this case, two control nuts can be used to mount the central controls far enough behind the panel in order to clear the receptacle contacts.

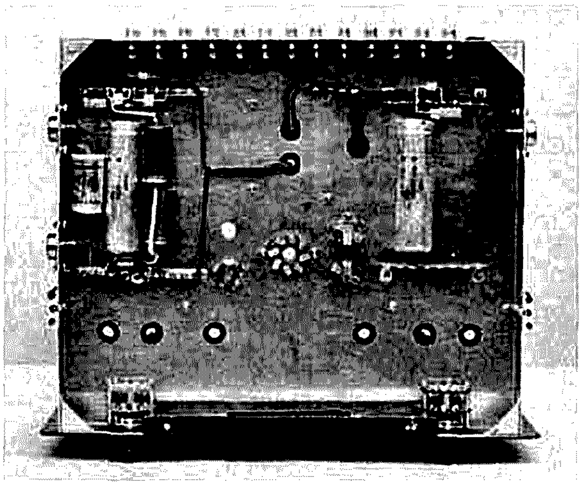

fig. 5. Bottom view of digital accessory showing power supply and unwired switches and other controls.

More holes for controls are shown in fig. 3 than were planned for expected requirements. Unused holes can be filled with snap-in plugs until required by future expansion.

read-out window

A rectangular hole is required in the panel for the read-out display. This can be slightly smaller than IC-type read-outs. The old way of drilling a series of small holes, then connecting them by one of several means, is a lot of work. There are two easier ways.

First, lay out the desired rectangle on the back side of the panel, marking all four sides. Drill small holes in all four corners just inside the rectangle. Then back away from the corners and drill large holes inside of the rectangle. If a neighbor has home shop equipment with a milling device, a small milling tool can be used to rout out the metal within the rectangle.

Without this, borrow a sabre (scroll) saw with a fine-tooth nonferrous metal-cutting saw blade. These saws are priced from under $10 at Sears and about twice that in the Skil tool line. Cloth should be placed on each side of the panel, then

wood blocks, before putting it in a vise for sawing. The rectangle can be cut by sawing in different directions from a starter hole, even by using the side of the saw blade as a file for thin cuts.

After the panel work is finished, hang the panel by a wire or cord through a screw hole in the bent-over angle lip (or through both holes), and spray with Krylon all-purpose gray (no. 1318) or a suitable alternate.

The same type of cutting operation can be used for the rectangular hole for the contacts of an R-644 Vector *plugbord* receptacle on the front of the chassis below the rectangular window in the panel.

Later, a Vector 3662 *plugbord* ($6.95), with P pattern 0.1-inch hole spacing, can have two inches sawed off the end of the board, 4-7/16 inches from the plug end, to retain 36 rows of holes including the two rows connected to the 44 contacts. This board will mount up to eight read-out units such as Minitron or light-emitting diodes, plus their decoder/drivers, and considerably more DIP ICs if desired.

The drivers should be mounted with the read-outs in order to have only four data leads per decade, not seven, to keep within the limit of 44 contacts. Boards with more contacts are available, but may not be necessary.

Normally, mounting the SN7475N quadruple bistable latches on this board is not desirable, because latches are to be omitted from the circuit when time is being displayed. Of course latches for the digits not required to display time can be mounted on this board to use up some of the available board space here, and relieve the other circuit boards which will be used for the time base, and for the counting functions.

If light-emitting diodes are used as read-out displays the voltage dropping resistors required between the decoder/drivers and the read-outs should be on this board. There must be sufficient contacts to handle decimal points. With eight digits displayed, there may not be sufficient contacts for future expansion.

hardware

Most of the circuitry is on Vector *plugbords* using the 3662 mentioned above for linear rf uses and the 3682-2 ($8.49) which has convenient ground and + V_{CC} busses printed on it for nonlinear ICs and the input circuitry.

These boards mount in the R-644 receptacles as shown in fig. 6. A convenient means of mounting is available by obtaining two 24-inch Vector TS240

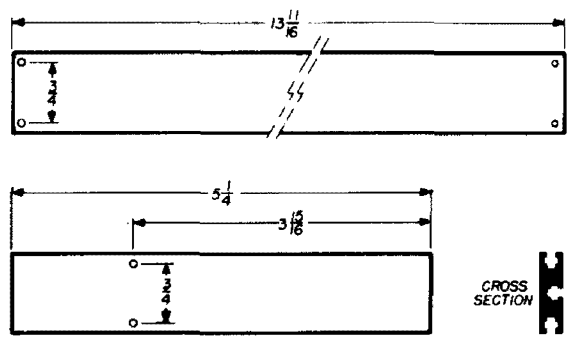

fig. 7. T-Strut lengths and locations of holes. Two pieces of Vector no. TS-240 24-inch stock are required.

T-struts.* The company continues to use this address although the earthquake caused them to move to another area.

Also, order a package of 24 NT6-3 plastic inserts, with no. 6 sheet-metal screws, SC6-11; and a package of 25 square 4-40 nuts, NT4-7. The screws and inserts make right-angle butt joints possible, and the square nuts are used to secure to chassis by the center T-slot nut cavity, and to fasten R-644 receptacles by one of the two edge T-slot nut cavities. (Small hex 4-40 nuts *will* work.)

Saw off, and file smooth, one top T-strut, one bottom T-strut, and two side T-struts as shown in fig. 7. Fig. 6 shows the side T-struts 3/8-inch longer than the dimension in fig. 7. This is satisfactory if the bottom chassis guide rails are wide enough to take the chassis plus both side

*Vector Electronics, 12460 Gladstone Avenue, Sylmar, California 91342, $1.50 each.

T-struts without more than minor springing, but it is a close fit.

It is necessary for the bottom strut to be shimmed up 0.1-inch above the chassis with two large washers so that the lower edge of the plugboards will be above the chassis surface. Some chassis space could be saved for pushbutton location by using two short pieces of strut in lieu of the long bottom one, and shimming their ends with washers held in place with screws through the chassis and up into nuts in the central T-slot nut cavity. This would result in a small loss in strength.

chassis holes

The more critical chassis holes are marked on **fig. 4.** Adequate 7/16-inch holes for large-hole grommets may be drilled *in front* of the lower T-strut and located convenient to the below-chassis control wiring. The holes for the crystal oscillator oven and clock reset pushbuttons were originally planned to be in front of the bottom T-strut. However, that results in possible conflict should nearby panel holes be used for longer switches or controls in the *future.*

The hole for an octal socket for the crystal oscillator oven should be punched or drilled with a fly-cutter; due to the thin, soft aluminum the ¼-inch fly-cutter center guide drill should pass through a ¼-inch bushing taken from an old switch or potentiometer, temporarily mounted to keep the drill from wandering.

Suitable holes for the transformers and power-supply parts, and for adequate pushbuttons for reset and adjusting the clock, should be drilled. Switchcraft type 103 are suitable spdt pushbuttons. The holes for the grounded and the insulated LM309K 5-volt voltage regulators can be drilled in the side aprons of the chassis to save top chassis space and provide adequate heat-sinking.

t-strut assembly

Prepare the lower T-strut (it has no holes) by inserting a plastic screw insert, properly oriented in the slot, into each side T-slot at one end of the lower T-strut, then cutting off the flat plastic that continues to extend *beyond the end* of the T-strut. Next, put a sufficient number of square 4-40 nuts, or small hex nuts if you do not have the Vector square ones, into the rear T-slot to take care of any future expansion. Eight or ten should be sufficient.

Now put plastic screw inserts into the other end of *the side T-slots, again* cutting off the flat surplus material. Put a 4-40 x 3/8-inch machine screw up through each of the four holes in the top of the chassis with a lock washer. On top of the chassis put on a shimming washer (not retained by the nut), and then a square nut, on the two screws that are clear of the sides. Before tightening the nuts, slide the lower T-strut onto these nuts using the central T-slot. Adjust the position at the sides and tighten the machine screws.

Next, place plastic screw inserts into the side T-slots at the top of each side piece of T-strut (away from the mounting holes), again removing excess plastic. Put 4-40 x ¼-inch machine screws, lock washers *and a square nut* through the four holes provided, from inside the chassis, using the mounting holes. Slide the side T-struts down over the square nuts, using the central T-slots, so that the side T-struts stand up outside of the sides of the chassis. Secure the machine screws. Insert no.-6 round-head sheet metal screws through the T-strut mounting holes drilled through the side T-slots and into the plastic inserts in the ends of the lower T-strut.

Now, using no.-6 flat-head sheet-metal screws, secure the top T-strut to the top of one side T-strut. Then insert the same number of square nuts into the rear T-slot of the top T-strut as were put into the bottom T-strut, and secure the remaining end of the top T-strut to the vertical side T-strut. Later, the R-644 receptacles can be attached – probably after soldering wires to the contacts. The completed assembly appears in **fig. 6.**

After the R-644 receptacles are attached it may be necessary to bevel

with a file the lower side edges of the Vector 3682-2 *plugbords* in order to ensure that there will not be a short circuit from the edge printed buss to the chassis (though both may be at ground potential). Later, it may prove to be advisable to place a strip of plastic tape

power supply

Inasmuch as the provisions for digital clock, latches, decoder/drivers, read-outs and digital dial mixers may consume considerable power, adequate amperage has been provided in the design. The most

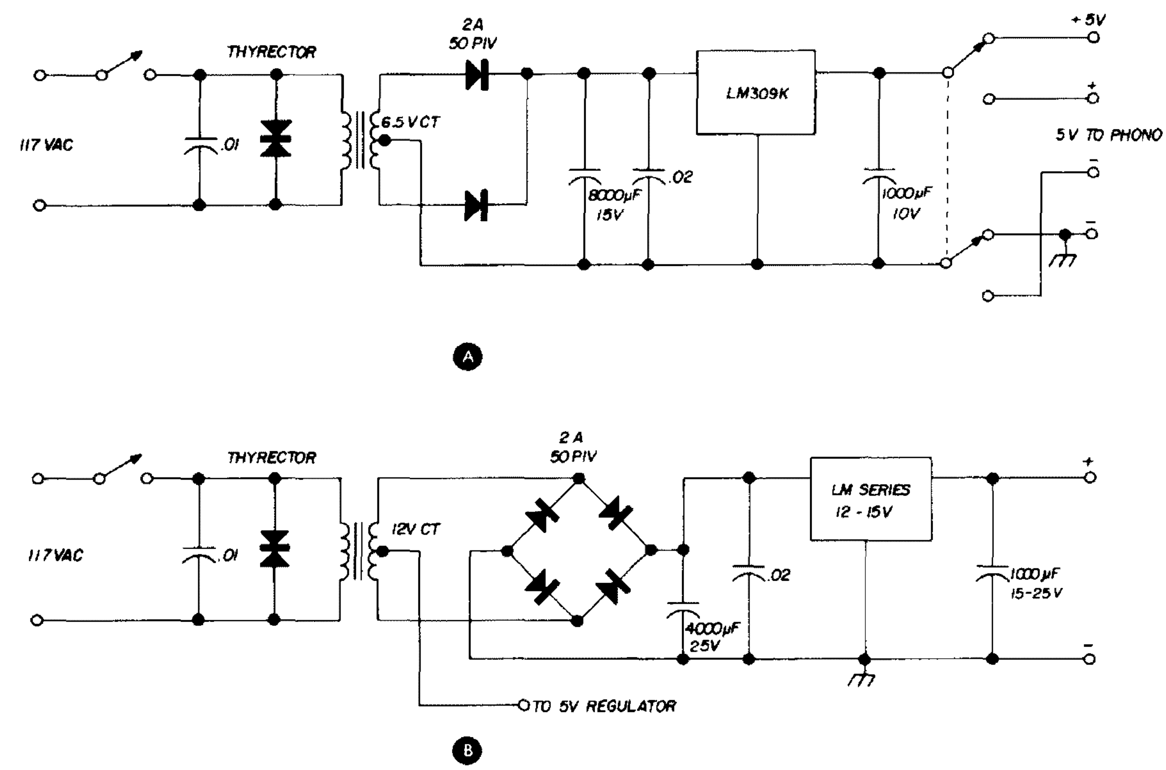

fig. 8. Power supply options. Center-tapped transformer, LM309K five-volt regulator and switched ungrounded output for experimental connection elsewhere is shown in (A). 12- or 15-volt supply using LM336 or LM337 regulator, with centertap to five-volt filter and regulator in (B).

along the top of the chassis where the plugboard will be, or to place tape around the bottom edge of the plugboard.

The rear view, fig. **6**, shows the far plugboard with its printed $+V_{cc}$ and ground busses extending out from the receptacle by use of a Vector 3690 card extender. This permits operation of a board in the clear for maintenance or test purposes. In a few cases the extension may change the operating characteristics of the board somewhat. An alternate method is to make up a cable extender using the R-644 receptacle and a 3662 board connected by 44 wires. The wires can be flat ribbon cable, but it is difficult to obtain this in less than 100-foot lengths.

convenient 5-volt regulator with internal overload protection appears to be the National Semiconductor LM309K for currents around one ampere. These will be seen in figs. **5** and **6** mounted on the chassis skirt to conserve chassis space.

The LM309K is in a diamond-shaped TO-3 case which requires heat-sinking if its full capability is to be used. This regulator has only an input, an output and a ground (negative) connection. It does not provide any other voltage, although there is a way to vary it by lifting the ground voltage with respect to the output. It is recommended that silicone heat-conducting heat-sink compound be used under the regulators, on both sides of the mica insulator.

Apparently the LM309K was designed to have an input voltage of 7 to 25 volts. Avoid loading the transformer, rectifier and input capacitor to a lower input voltage to the regulator. This is provided by a good 6.3-volt transformer; with a large filter capacitor of some 8000 μF for the 5-volt supply, and half that for 10 to 15 volts. A good 2-ampere transformer for under $4 is the Calectro DI-747 (GC Electronics), producing over 9 volts input to the regulator from a 12-volt center-tapped rectifier, before a load is applied. The use of a 12-volt center-tapped winding permits a bridge rectifier for the higher voltage, and half that voltage from the center tap, as shown in **fig. 8.**

higher voltages

National Semiconductor also produces the LM336 for 12 volts at about 500 mA, and the LM337 for 15 volts at about 400 mA, depending upon both the input voltage (which should not be excessive) and the adequacy of the heat sink. With one of these regulators a center-tapped transformer of sufficient voltage can be used simultaneously for the 12- or 15-volt regulator, and also for the 5-volt regulator.

For that matter, it is feasible to use two regulators or more for different voltages from the same higher-voltage rectifier, provided that the maximum regulator voltage and heat dissipation ratings are not exceeded. The higher-voltage regulators are priced around $6. All three of these ICs are protected by internal current-limiting.

With any of these power supplies rectifier diodes that handle 1.5 to 3 amperes may prove convenient. Glass-amp units can break unless all bends in the leads are made between two long-nosed pliers so that there is no strain on the glass.

For current less than about 45 mA the adjustable type 723 IC voltage regulator (made by Fairchild and others) may be used. It is rated up to 150 mA, which would require a 0.335-inch clip-on heat sink if the metal package is selected. The total internal power dissipation must be limited to 800 milliwatts. The input voltage must not exceed 40 volts. There must be at least a 3-volt drop between input and output voltages.

The current-limiting resistor, R_{sc}, can be selected by dividing 0.65 by the desired limiting current in amperes, and frequently is ten ohms (see fig. 9). Approximate values for the R1/R2 voltage divider are given in table 1. These figures are derived from a Fairchild application note with some adjustment to approach standard resistor values. R3 is needed only for minimum temperature drift; it can be calculated from R1 x R2/(R1 + R2), and frequently is around 3k ohms. **Fig. 9** shows a circuit used for an output above 7 volts.

Avoid blowing out the 723 IC by accidental application of reversed rectifier voltage. This can be done by using a series diode between the rectifier and the IC.

Circuits, including those for external-pass transistors, will be found in the Fairchild application note for cases where more than the rated current is to be regulated by the 723, or the output voltage is to be below 7 volts.

Keep in mind the much higher peak input voltage to the regulator. A 15-volt capacitor or larger is required for the first filter capacitor at the output of the 5-volt rectifier diodes, and a 25-volt or larger capacitor in the 12-volt supply. A lower break-down voltage capacitor is satisfactory for the output side of the voltage regulators, just exceeding this regulated voltage. Generally, a smaller capacitance is suitable on the output of the regulator, since additional capacitors generally are placed on the circuit boards because of the tendency of TTL flip-flops to create spikes and noise which can trigger other flip-flops.

If mica mounting kits are used to insulate the TO-3 regulator cans, be sure that there are no burrs around any holes, and test the unit for high-resistance isolation from chassis or heat-sink with an

ohmmeter before connecting it to other parts.

In the unit pictured in figs. 5 and 6 one 5-volt power supply is completely insulated from ground until it is switched into the circuitry, so a pair of phono jacks can bring the voltage out to other devices, such as the Palomar Engineers' electronic keyer which requires a grounded positive. Other phono jacks bring out the grounded 5- and 12-volt supplies for other uses.

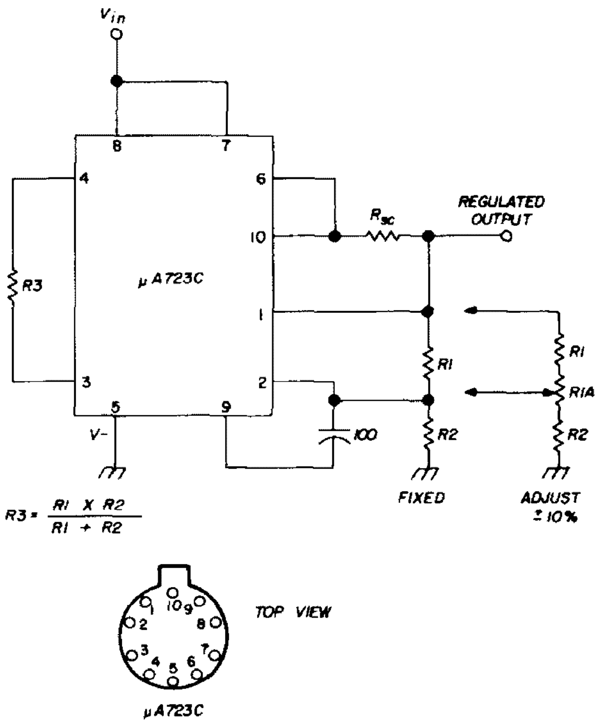

fig. 9. Fairchild μA723-type IC precision voltage regulator circuit for use above seven volts, without series-pass transistor. Input must exceed output by at least three volts.

frequency standard

Initially you can use the 60-Hz power (or 50-Hz elsewhere with three flip-flops in a divide-by-five circuit, or the BCD section of a 7490 decade[1]) for satisfactory operation of a digital clock and for equipment digital dials for tenth-kilohertz read-out. This was discussed in *ham radio*[2] in connection with measurements made from a line-frequency-controlled counter using Fairchild 9093 DTL dual JK flip-flops.

The 9093 flip-flops have toggled on ac at all frequencies from a few hertz up to their limit of 2.2 MHz without any input

table 1. Approximate values for the R1/R2 voltage divider in fig. 9.

output voltage	fixed output R1	fixed output R2	adjustable output R1	adjustable output R1A	±10% R2
9	1800	6800	750	1k	2700
12	4700	6800	2000	1k	3000
15	8200	7500	3300	1k	3000
28	22000	7500	5600	1k	3000

circuitry. Three FFs can divide by two and three, and four FFs (or a decade counter) can divide by ten, to give a one-second frequency. The remaining FF in the dual units can produce the one-second on and one-second off frequency desired for counter gates.

The 9093 and some other FFs can be driven directly by a volt or so of ac. Some may require a diode across the input to pass the negative half of the cycle. Usually, several volts of ac are available between ground and a terminal in a power supply, even from a bridge rectifier without a transformer center-tap. A stopping capacitor can be added; then, if desired, a 2-volt zener or several series silicon diodes can be added with anode grounded to provide the FF connected across the zener or diodes with a volt or two of ac safely. This also eliminates the unnecessary negative excursion of the ac sine wave.

Another source of ac is the power line itself, through a nonpolarized capacitor, then the resistor and zener[3] mentioned above. With a power-line source the accuracy of a counter will be within about 0.02 percent on any count, and the average of a number of measurements will probably be correct.

Be sure to provide an octal socket hole in the chassis for future addition of an oven-controlled crystal oscillator. The common, inexpensive E-cut 100-kHz crystal is not nearly as good as a DT-cut crystal for 100 kHz or 1 MHz. The DT-cut has a broad flat temperature curve.[4] The cost is a few dollars more. For best results, the crystal and oscillator should be in an oven.

The counters I have use 100-kHz and 1-MHz oscillator ovens made by Monitor

Products.* One has a 400-kHz crystal, IC flip-flop oscillator and two frequency-dividers which use cross-connected gates. The 100-kHz output is suitable to drive DTL and TTL dividers with fair isolation. However, if the output is taken to other devices it would be well to connect these and the time-base dividers to the oven output through isolating gates.

The oven operates on 120 volts ac, though proportional control would require at least pulsating dc. The oscillator operates on 10 to 12 volts, being zener-regulated internally to 5 volts. A 20-percent change in voltage causes about a 3-Hz change in the 14-MHz harmonic, so it is well to feed the oscillator from a regulated power supply or, at least, from a stable power supply that does not feed a varying load.

The oven is left on except during foreign trips; no adjustment in crystal frequency appears to be needed except possibly once a month, or for an ARRL frequency-measuring test in which an error of one or two Hertz may be undesirable.

C. & H Sales* has sold a large number of James Knights (now CTS Knights), oven/oscillators, 1 MHz, model JKO-PIP-X96D, with proportional ovens. These ovens also plug into octal sockets but operate from 23 to 36 volts dc, both for the oscillator and the oven. Varying the voltage on the oscillator down to eight volts controls the frequency over a considerable range. A hole plug in the top can be unsoldered in order to reach the frequency-adjusting capacitor.

Probably there are other oscillator/oven sources, both surplus and new, such as Bliley and CTS Knights. Only mercury thermostats and proportional control will be free of occasional clicks in a receiver.

*Monitor Products Company, 815 Fremont, South Pasadena, California 91030.

*C & H Sales, 2176 East Colorado, Pasadena, California 91107.

*The SN74162N synchronous up/down decade counter is available from Polypaks, Box 942, South Lynnfield, Massachusetts 01940.

up/down counters

Before concluding it may be desirable to make preliminary mention of the SN74162N synchronous up/down decade counter, which is fully programable[5] – that is, it can start counting at any number, and even divide by odd figures like seven. This has become available under $5 and need be used only in decades that are displayed on the read-out.*

For a year or more these devices have provided a newer approach to decade frequency division; like the 7490 decade they divide the frequency, but they also provide a down-counted read-out. There is good reason to mention these because of a possible simplification of equipment by eliminating all circuitry used for synthesis of the original signal frequency before counting it.

reading material

Preliminary reading material on non-linear ICs, counters and clocks will be found in the following reference material.

references

1. R. Wilson, G3TBS, "A 24-Hour Digital Clock," *Radio Communications,* April, 1971, page 254.
2. E. H. Conklin, K6KA, "Counter Gating Sources," *ham radio,* November, 1970, page 48.
3. D. Kochen, K3SVC, "Transformerless Power Supplies," *73,* September, 1971, page 14.
4. I. M. Hoff, W6FFC, "The Mainline FS-1 Secondary Frequency Standard," *QST,* November, 1968, page 34.
5. E. H. Conklin, K6KA, "Electronic Counter Dials," *ham radio,* September, 1970, page 44.
6. H. S. Knoll, WAØG0Z, "Simplified Digital-Counter Readout," *ham radio,* January, 1970, page 66.
7. O. C. Stafford, "Electronic Digital Clock," *ham radio,* April, 1970, page 51.
8. A. A. Kelley, K4EEU, "Digital Frequency Counter," *ham radio,* July, 1970, page 16.
9. A. A. Kelley, K4EEU, "10:1 Digital Frequency Scaler," *ham radio,* August, 1970, page 26.
10. R. M. Vaceluke, W9SEK, "Modular IC Counter Circuits," *ham radio,* August, 1970, page 63.

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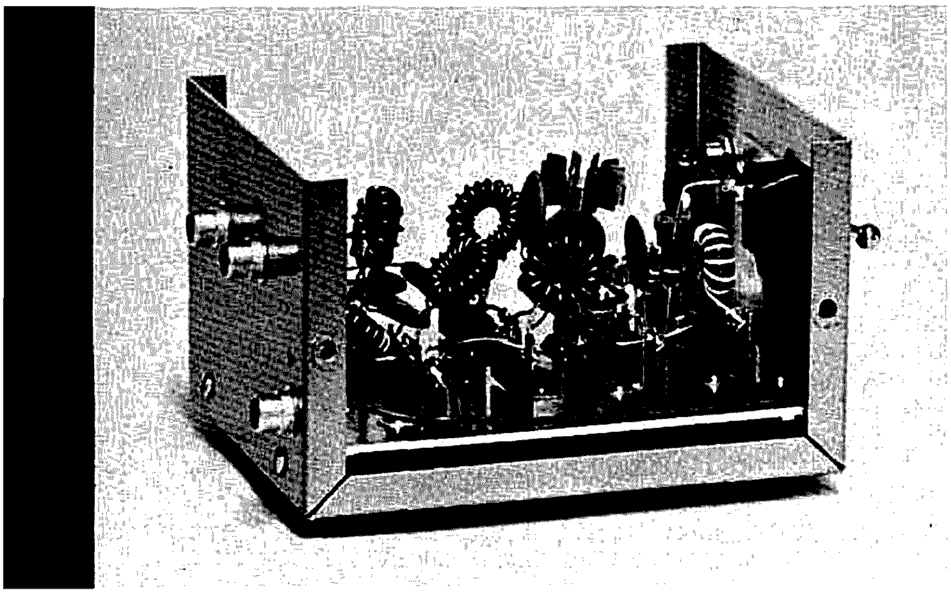


solid-state driver

and final for 40 and 80 meters

A project for one or two weekends — features include antiresonant traps, making band switching unnecessary

C. E. Galbreath, W3QBO, 8326 Still Spring Court, Bethesda, Maryland

This two-band solid-state transmitter was designed for use with a vfo such as the one described in an earlier issue of *ham radio.*[1,2] It operates on 40 and 80 meters, consists of only two stages, and incorporates the best design practices for stabilization and unwanted signal suppression. The unit is easy and inexpensive to build and is an excellent performer. Output can be varied from a few milliwatts to over two watts.

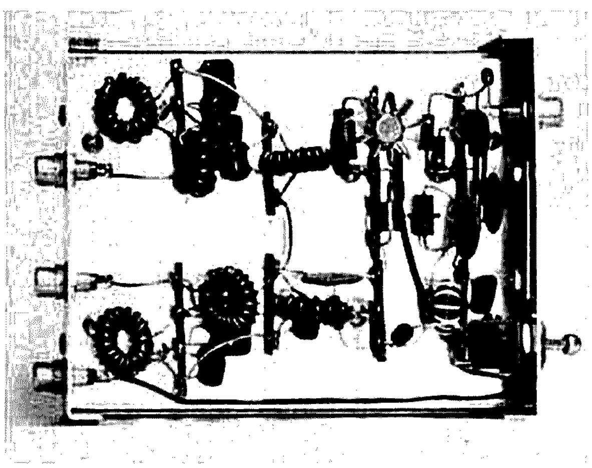
the circuit

The driver stage (fig. 1) uses an inexpensive Motorola 2N4124, Q1, which operates as a class-B amplifier with its base biased through a voltage-divider to obtain approximately collector-current cutoff. With no signal applied collector current is near zero, which minimizes current drain during key-up conditions. As a class-B amplifier, the stage is easier to drive than if operated in class C. The signal input from the vfo is coupled to the base of Q1 at low impedance. A 100-ohm resistor in series

with the input at the base of Q1 eliminates any tendency to self-oscillation, which sometimes may occur without it. The signal input at the base of Q1 is not appreciably reduced by this resistor.

Other transistors may be substituted for Q1; for example, I've used a Motorola 2N3904 with good results. No heat sink is required for Q1.

The driver tank circuit includes a broadband toroidal inductor, L1, which is



Wiring of the 40-80 meter transmitter.

fixed-tuned by C1. L1 resonates on both bands without changing the tank-circuit capacitance. The emitter-bias resistor of Q1 is unbypassed to provide a small amount of degenerative feedback, which increases driver stability. Interstage coupling through the B+ line is reduced by the 25 μH choke between driver and amplifier. Added protection is provided by the .05 μF bypass capacitor between B+ and the chassis.

The driver output is inductively coupled to the amplifier transistor through L2, which is wound over L1. L2 also provides the correct impedance match for the excitation voltage to the amplifier.

output stage

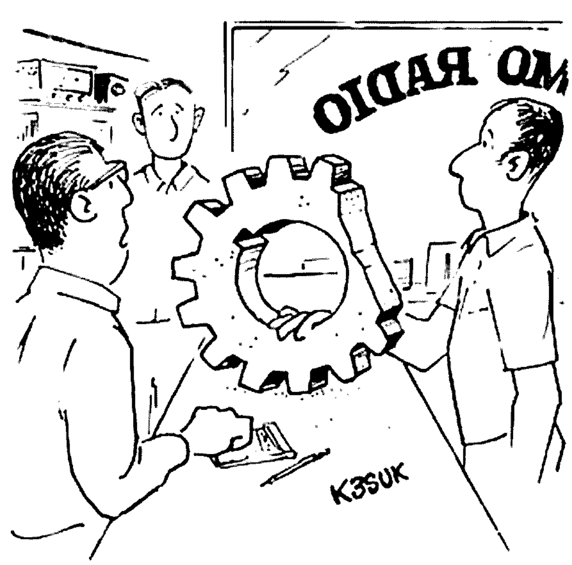
The final amplifier consists of a 2N2102 transistor, Q2, whose tank circuit is a double pi network for each band. Toroidal inductances are used in the pi networks as they are self-shielding and occupy little space. They are also easy to make and mount in the circuit. The output impedance of the final stage is approximately 50 ohms. A heat sink is required for Q2.

No switching is needed to change bands. An antiresonant trap is inserted ahead of each pi network. On 80 meters the trap ahead of the 40 meter pi network is resonant at 80 meters and isolates the 40-meter tank from the output circuit. On 40 meters the trap ahead of the 80-meter pi network is resonant at 40 meters and isolates the 80-meter tank from the output circuit. I've used traps for this purpose since early 1968.* When changing bands, it's necessary only to insert the transmission-line plug into the appropriate output jack at the rear of the Minibox.

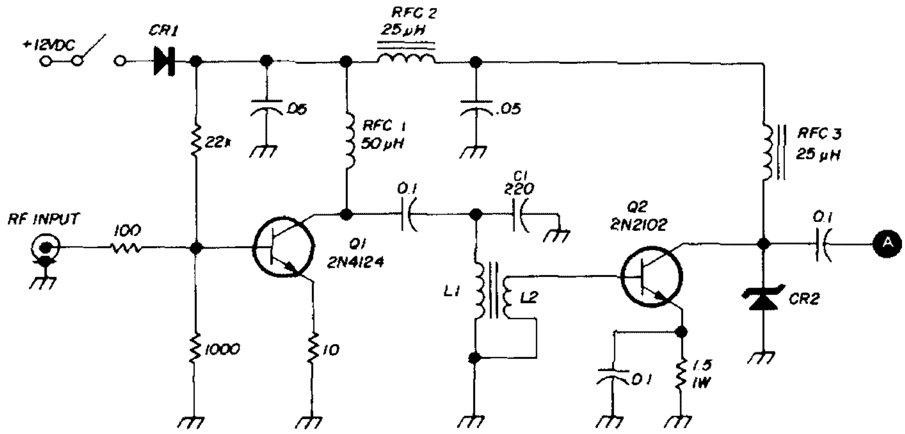
performance

The rig runs about 2 watts output using a 12 V supply. The two stages together draw about 320 mA. The voltage drop across polarity-guarding diode CR1 is over 1 volt. The transmitter will operate on as little as 3 volts, reducing the output to a few milliwatts. For regular duty, a 12-volt lantern battery makes a good power supply. A battery pack of 8 or 10

*They were used in 1967 in a transmitter known as the ARP-1.



"Joe, we got any parts for those 160-meter beam rotators we were selling last year?"



C1	220 pF dipped silver mica capacitor	L1	24 turns No. 24 enamelled on Amidon T-50-2
CR1	50 volts, 1.5 amps	L2	12 turns No. 24 insulated over L1
CR2	39-volt, 1 watt zener, American Semiconductor Corp. 1N3034 or equivalent		

fig. 1. Schematic of the 40-80 meter transmitter. Traps ahead of each output network eliminate bandswitching.

D cells in series will also do very well. For reduced power a 6-volt lantern battery or a 6-volt battery pack will work well. Although I have operated the transmitter on 15 volts, I have not established a safe maximum B+ voltage.† A 100 μF electrolytic capacitor across the battery terminals provides an ac path to ground.

A zener diode, CR1, between the collector of Q2 and ground protects the transmitter from spikes in excess of 36 volts. Too much voltage on the collector will destroy the transistor. Normally the peak rf swing on the collector will not exceed twice the supply voltage.

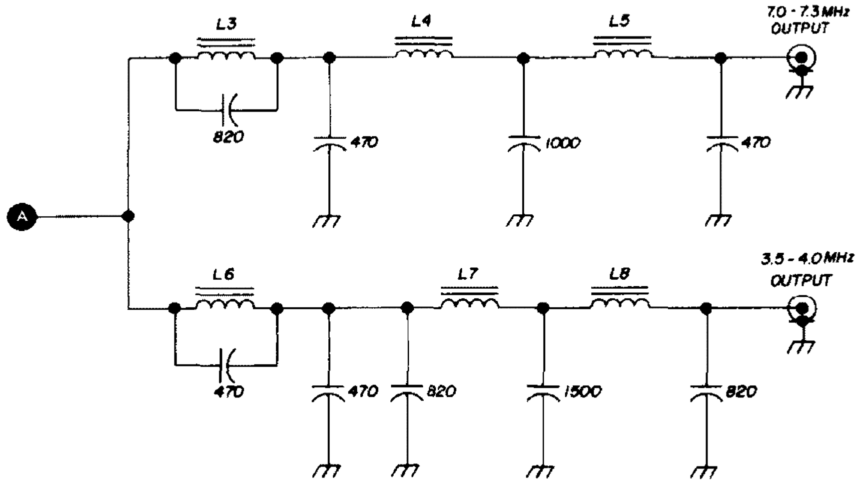
construction

The transmitter is housed in a 3 x 4 x 5-inch Minibox. Components are mounted on terminal strips for ease of construction. The terminal strips may be mounted either on the base of the Minibox or on an aluminum plate, which may be inserted into the box after component wiring is completed. I prefer the second method, as it simplifies component mounting and wiring, and the unit can be tested before inserting it into the box. The aluminum plate is bolted to ½-inch right-angle brackets at each end of the Minibox.

Start construction with the angle brackets. I used ½-inch right-angle aluminum purchased at the local hardware store. Cut two brackets about 3½ inches long and mount one at each end of the Minibox. Cut a piece of aluminum 3½ x 4-7/8 inches and bolt on the brackets in the box. Now remove this piece with the brackets attached and use it outside the box as the chassis on which to mount the terminal strips and assemble and wire all components.

The driver stage is located at the front end. All components for the driver stage except the tank circuit are mounted on a 5-lug terminal strip, center lug grounded. The driver-stage tank is mounted on a 3-lug terminal strip in line with the 5-lug strip. Similarly, all components for the final except the tanks are mounted on a 5-lug terminal strip. A 3-lug strip in line with this 5-lug strip holds the B+ line decoupling choke and bypass capacitors. Q1 and Q2 may be soldered directly into the circuit, or transistor sockets may be used with two of the socket terminals in

†Motorola's Semiconductor Data book lists maximum V_{cb} ratings of 30 V (2N4124) and 120 V (2N2102). *editor.*



L3,L7,L8	18 turns No. 20 enamelled on Amldon T-68-2	Q2	Motorola 2N2102 (heat sink required)
L4,L5,L6	13 turns No. 20 enamelled on Amidon T-68-2	RFC1	50 μH rf choke, Millen 34300-50
Q1	Motorola 2N4124	RFC2,RFC3	25 μH rf choke, Millen J-300-25

each case soldered firmly to two lugs on the terminal strips. Placement of resistors and capacitors on the terminal strips is not at all critical. The use of terminal strips assures ease of mounting and short direct connections.

Each trap is mounted on a 3-lug terminal strip, center lug grounded. Each double pi-network tank is mounted on a 4-lug terminal strip, one lug grounded.

table 1. Typical rms signal voltages.

	80 meters	40 meters
output of vfo	4.0	4.4
at base of Q1	3.0	3.4
output of Q1	9.2	8.0
at base of Q2	2.7	2.9
output of Q2	15.0	8.0
final output (tank)	10.0	9.0

The two tanks with their associated traps are positioned on either side of the centerline toward the rear of the chassis.

After all components have been mounted on the chassis and wiring completed, check the circuit, point to point, for possible wiring errors. Operation of the two stages can then be tested before mounting the chassis in the Minibox. Use a two-watt, 50-ohm composition resistor as a dummy load. A sensitive swr bridge and reflected-power meter connected between transmitter and dummy load is most useful for checking operation. The swr should be 1:1; there should be no reflected power indication on the meter. Meter deflection in the forward direction indicates transmitter output. Those with a vtvm and an rf probe can check rf voltages at various points in the circuit. Table 1 gives typical rms rf voltage readings as measured by a Heath vtvm and a Heath rf probe when using a 12-volt power supply. Different transistors of the same type will often produce somewhat different results.

On the front of the Minibox is the on-off switch for the power supply and the phono jack for the vfo line. On the rear of the box are 3 phono jacks, one for the output on each band and the third for the power-supply connection. The holes should be drilled and the switch and jacks fitted before the completed chassis is put back into the box. The switch and jacks, of course, must be removed to permit installation of the chassis.

For good results on the air be sure to use an efficient antenna system.

reference

1. C. E. Galbreath, W3QBO, "A VFO for Solid-State Transmitters," *ham radio,* August, 1970, p. 36.
2. C. E. Galbreath, W3QBO, "VFO Buffer Amplifier," *ham radio,* July, 1971, p. 66.

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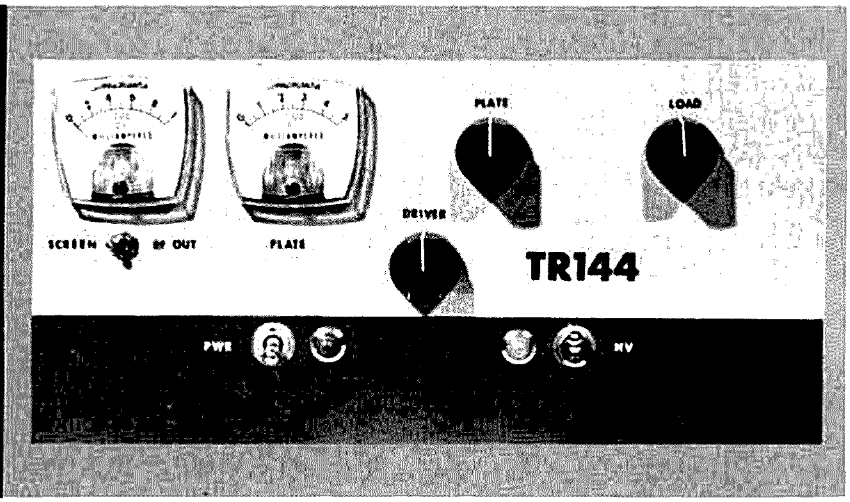


the TR-144

a transverter for two meters

Introducing another vhf transmitting converter and power amplifier compatible with Drake equipment — a companion to the TR-50

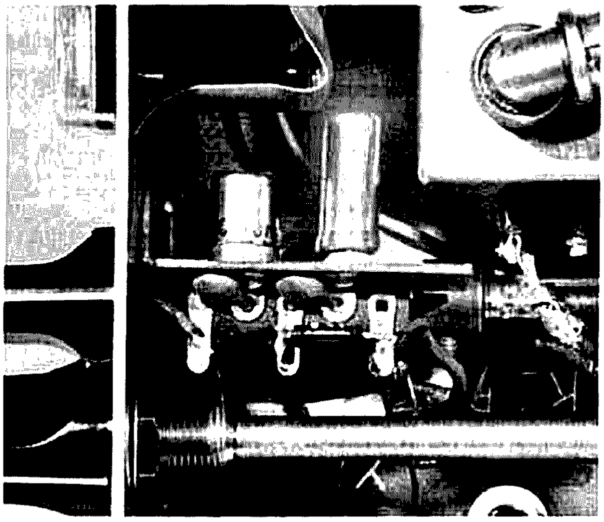
Louis E. Savoie, K1RAK, 29 Hillsdale Road, Holbrook, Massachusetts 02343

The TR-144 is a two-meter version of the TR-50 transverter described in an earlier issue of *ham radio.*[1] Like the TR-50, the TR-144 is compatible with Drake TR-3 equipment. Features include 350 watts PEP input (two-tone test signal), self-contained power supply, relay switching circuits, and a high-performance mosfet receiving converter. An Eimac 4X150 is used in the power amplifier. Full metering of critical circuits is provided. Additionally, the screen-current meter doubles as an rf-output indicator, which is desirable for initial tuning and on-the-air monitoring.

packaging

The cabinet is a TR-6 unit, purchased from Drake for $20. The chassis is made of .047 copper-plated steel (the plating isn't necessary; merely a preference). Chassis measurements are 10½ x 13 1/8 x 2 inches. The front panel, which is secured to the chassis by the switch and panel-light hardware, is .060 aluminum, 5½ x 10½ inches.* Captive hardware is pop-riveted to match mounting holes in the TR-6 cabinet.

*Full-scale drawings of the TR-144 front panel and chassis are available from *ham radio* for $.50 and a self-addressed stamped business envelope. *editor*



New oscillator circuit is built on a single-side PC board and mounted under the retaining hardware of the driver tune control.

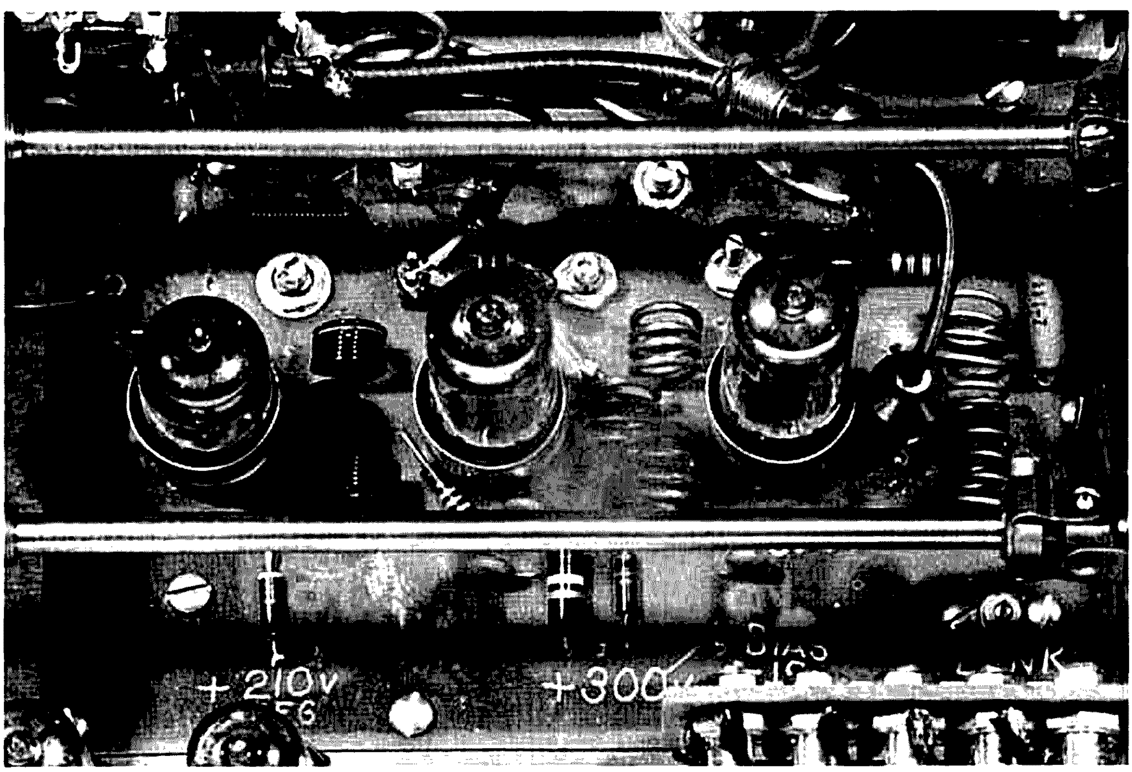
Much of the power and control circuitry had been designed for the TR-50 and is adaptable to the TR-144. The power transformer provides high and low B+ voltages, bias voltage, filament power, and relay-switching power. A series-phasing circuit provides the bias and relay voltages from the multiple secondary windings of the transformer.

The mixer (fig. 1) is the PC transmitting converter described previously.[2] It's the big brother of the six-meter mixer described in the same article, which used a 12AT7 oscillator feeding a 6360 doubler into a 6360 mixer-driver. (This circuit was changed for the TR-144, as shown later.) Initial bench tests of the mixer showed an output of 1½ watts with the voltages suggested. The mixer board requires no modification. However, some of the coils in the ocillator and buffer sections must be changed to accommodate the 28-MHz heterodyning system. Visual monitoring of the mixer output with a scope indicated a faithful reproduction of the Drake sideband generation.

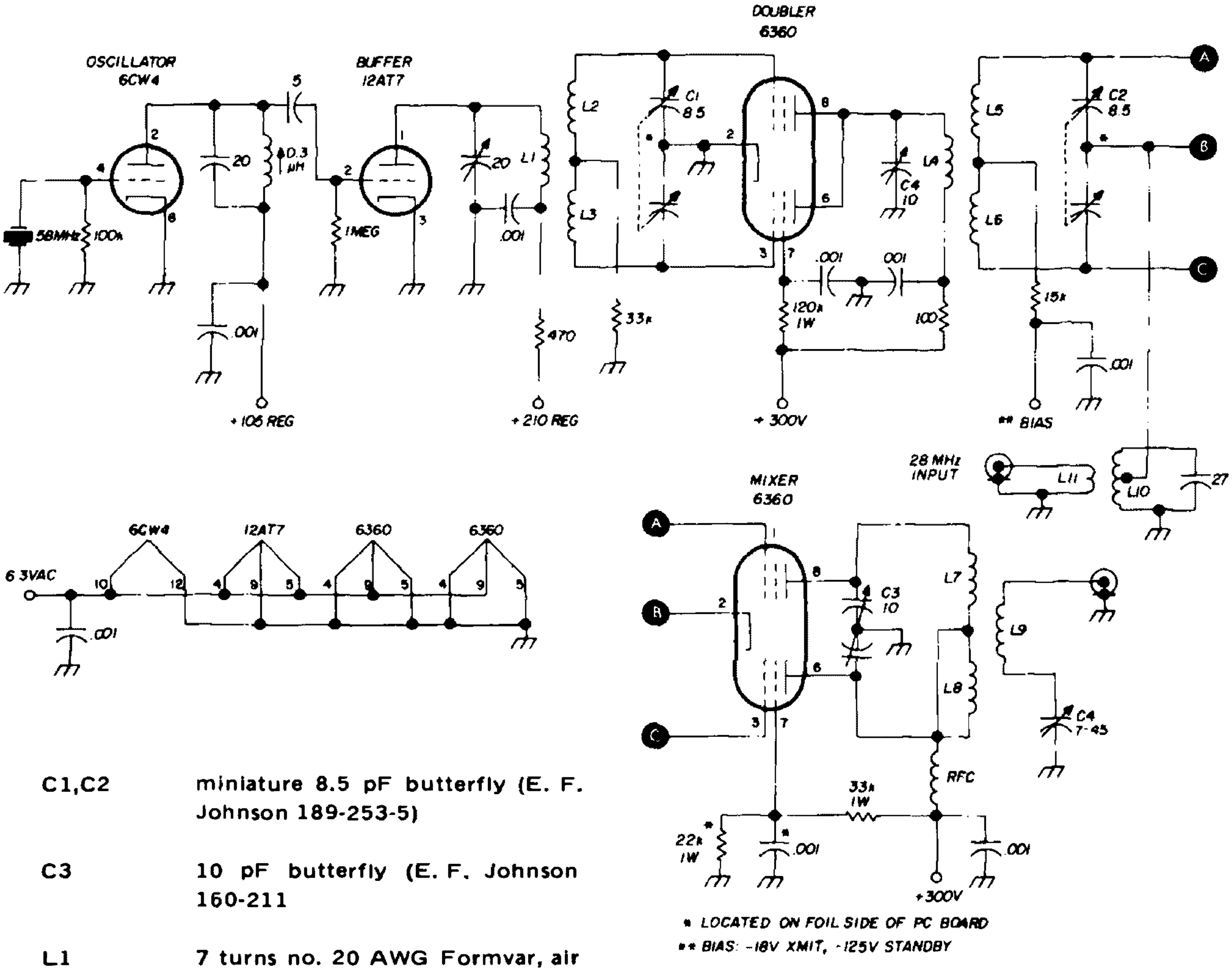
power amplifier

Based on previous experience and the success of the TR50, I decided to stay with the 4X150/4X250B power-tetrode family for the TR144 power amplifier. A push-pull circuit would yeild more output power, but limited cabinet space dictated a single-ended power stage.

The circuit (fig. 2) is straight-forward. Conventional vhf construction practices should be used. The power-amplifier plate circuit is enclosed in a 3 x 4 x 5-inch



Top view of transverter. A double flex shaft arrangement connects front-panel control to mixer output-tuning capacitor.



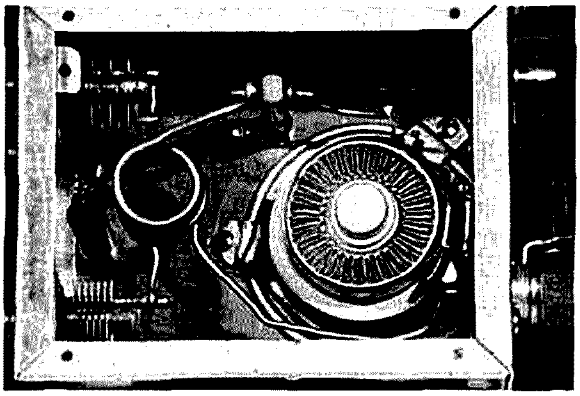
C1,C2	miniature 8.5 pF butterfly (E. F. Johnson 189-253-5)
C3	10 pF butterfly (E. F. Johnson 160-211
L1	7 turns no. 20 AWG Formvar, air wound, 3/8" dia., close spaced
L2,L3	9 turns no. 20 AWG Formvar, air wound, 3/8" dia., close spaced
L4,L5,L6	4 turns no. 16 AWG tinned copper wire, air wound, 3/8" dia.; turns spaced one-half wire diameter
L7,L8	4 turns no. 16 AWG tinned copper wire, air wound, 3/8" diameter; turns spaced 1 wire diameter
L9	2 turns no. 16 AWG tinned copper wire, air wound, 3/8" diameter; spaced 1 wire diameter
L10	12 turns no. 20 AWG Formvar, close wound on ¼"-diameter slug-tuned form, tapped 4 turns from cold end
L11	2 turns no. 20 AWG *on cold end* of L10
L12	8 turns no. 26 Formvar on ¼" dia. slug-tuned form
RFC	Ohmite Z-144
Xtal	58 MHz third-overtone

fig. 1. Modified two-meter transverter (from an original circuit by K1ISP in the April 1969 issue of ham radio).

aluminum utility box mounted on the chassis. Universal joints and ¼-inch steel shafts connect tuning controls to the front panel.

The grid-compartment box is made of 0.47 steel and mounted on the chassis underside flush with the rear-chassis panel. The 4X150 tube-socket hardware secures the box to the chassis undersurface.

The blower is a 15 cfm unit bracketed to the rear chassis wall and utility box. A



TR-144 plate-circuit compartment. Output link to antenna is mounted directly below plate coil. Plate-tuning capacitor is modified with alternate rotor and stator plates removed.

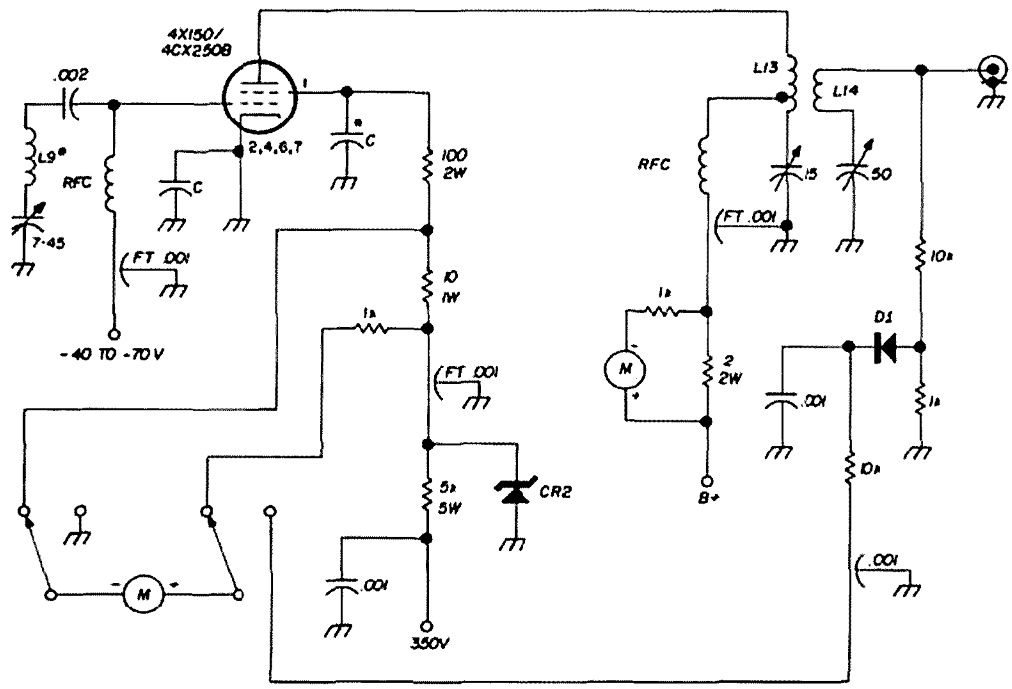
one-inch-diameter hole through chassis and grid box accepts the blower output. The mating edges of the blower outlet and chassis are sealed with a fillet of RTV silicone. The grid-box cover has a rubber gasket to inhibit air leakage.

Power- and filament-supply connections to the 4X150 are via .001-μF feedthrough capacitors. Rf coupling from the mixer link to the 4X150 grid is via an unbypassed ceramic feedthrough (photo). This arrangement provides excellent grid/plate circuit isolation, and the amplifier requires no neutralization. Should instability occur, conventional stub neutralization should stabilize the circuit.

power supply

Power for the TR-144 is supplied by a single transformer (fig. 3).* A full-wave rectifier/filter provides 900 Vdc (key down) for the amplifier B+. Twelve Vdc for the receiving converter and relay-switching circuits are obtained from

fig. 2. Power amplifier circuit. Metering scheme features switching circuit to monitor rf output.



C*	.001 μF 600 V disc from each cathode pin to ground screen bypass (part of Elmac SK600 socket)
L9*	link on mixer board
L13	4 turns no. 10 5/8 in. dia., 1 in. long. Tap 1 turn from bottom end
L14	2 turns no. 12 5/8 in. dia. spaced 1 wire dia. Mount ¼ inch from cold end of L13
RFC	Ohmite Z144
D1	1N34
CR2	1N2990B (ten 33V 10 W zeners in series)
M	Micronta 0-1 mA (Radio Shack, $2.98)

*The transformer is available from J. Reeves, WA9HKE, 2207 Columbus Ave., Anderson, Indiana 46014. Ask for P&H transformer.

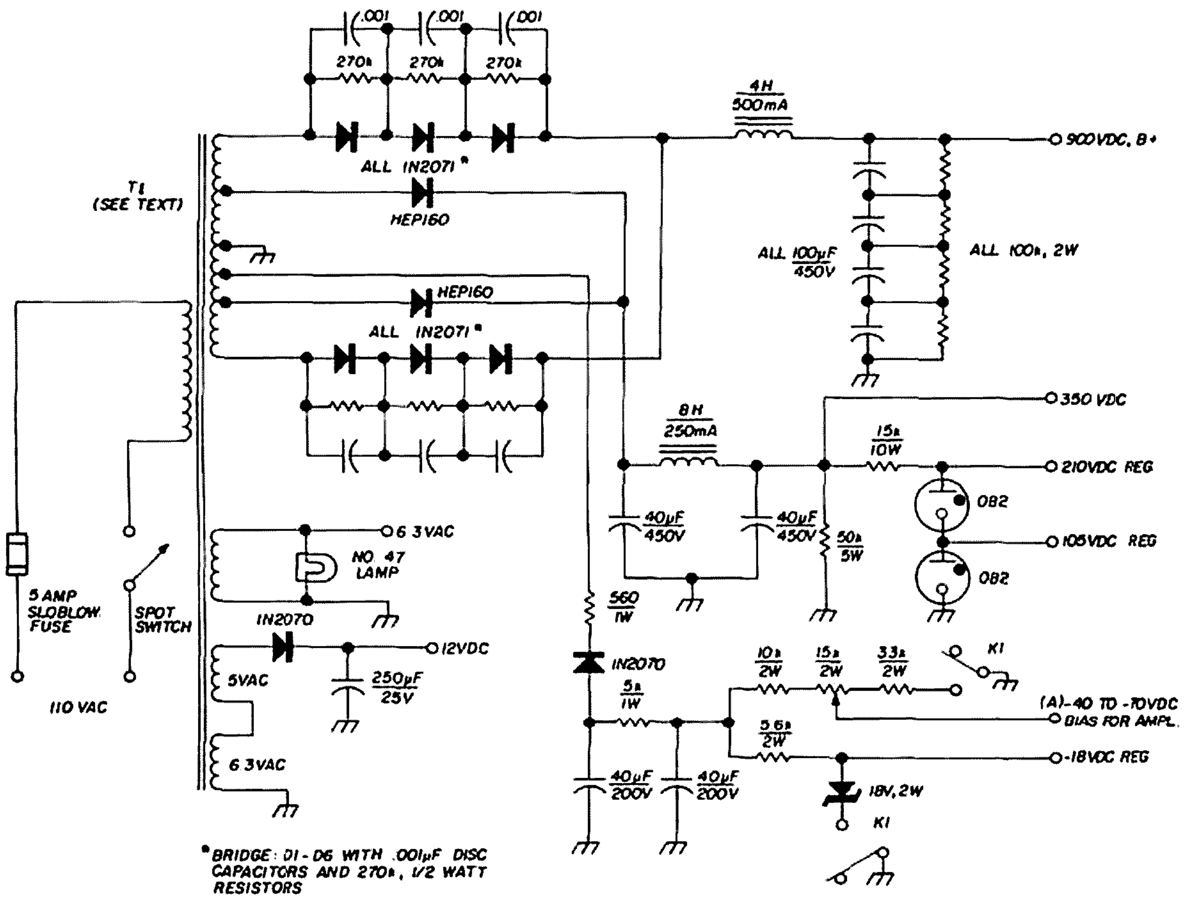
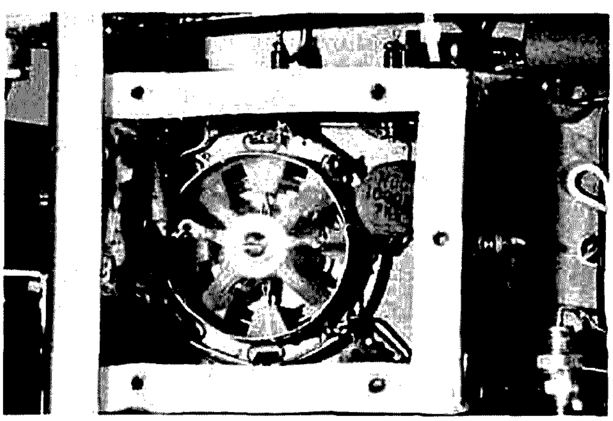


fig. 3. TR-144 power supply. High and low voltages for all circuits are provided by a single power transformer.

series-connected 6.3- and 5-Vac secondaries. Voltages for the transverter stages are obtained from taps on the HV secondary.

Ten series-connected zeners provide regulated screen voltage for the 4X150. Any combination of zeners that will deliver 350 Vdc and dissipate 20 watts is suitable.



Amplifier grid compartment. Filament screen, and bias voltages are supplied via feedthrough capacitors. Input from the mixer is via unbypassed ceramic feedthrough connectors mounted on front of box.

metering

The TR-144 metering scheme is the same as that used for the TR-50. Two 0-1 dc milliammeters in series with 1k-ohm ½-watt resistors monitor screen and plate current. Shunts for the screen- and plate-current meters (0-100 and 0-500 mA) are respectively 10 ohms 1 watt and 2 ohms 2 watts. The screen-current meter may be switched to indicate rf output.

Numbers on the meter scales were removed with a typewriter eraser and replaced with press-on transfers to indicate the desired ranges. Each meter face was sprayed with two light coats of clear acrylic lacquer to restore gloss and secure the transfer numbers.

switching and controls

The relay controls are identical to those in the TR-50 (fig. 4). The 10-dB pad, which must be used to reduce ssb exciter output, is switched into the system in the transmit mode.

All rf connections between the TR-3, TR-144, and 10-dB pad are via short lengths of RG-58/U cable and BNC connectors. The main control relay, K3, is a 110-Vac unit by Potter-Brumfield, type KA4314-1, which is actuated by the keying line from the TR-3. Relay K3, in turn, actuates the Dow-Key antenna changeover relay, K1, and the miniature coax relay, K2 (a surplus item). Relay K2 switches the TR-3 to the 10-dB pad on transmit. Relay timing causes no problem, and VOX action is good.

receiving converter

The receiving converter, available from Spectrum International,* is model DGTC 22 and features mosfet rf and mixer stages. It has a gain of 25 dB and is designed for 28-32 MHz output.

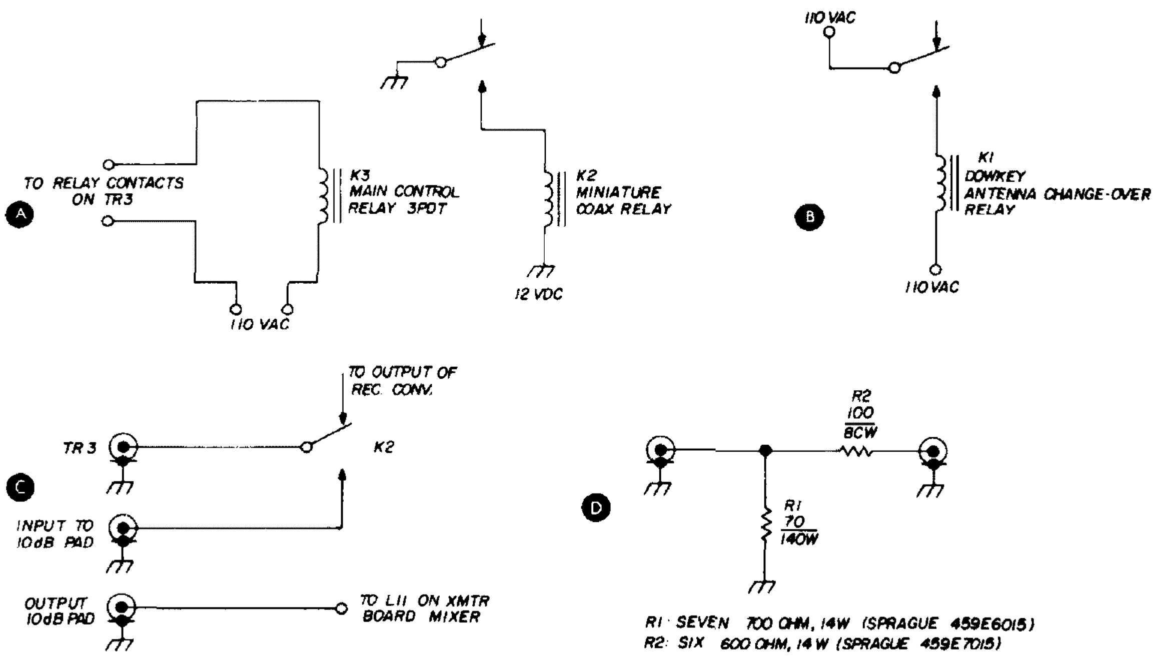
The converter is mounted in a 4 x 2 1/8 x 2¼-inch Minibox. The box cover has BNC fittings and feedthroughs for 12 Vdc and oscillator input. The feedthrough for oscillator injection is an unbypassed type.

The converter must be modifed slightly for transverter operation. The converter oscillator was disabled and the 0.5-pF coupling capacitor removed. It took some looking to find this capacitor in the converter. It turned out to be nothing more than a small loop of wire mounted next to the coil in gate 1 of the 3N140 mixer mosfet.

Remove the cover can of this coil by carefully unsoldering its tabs from the PC board. Unsolder and discard the wire loop, then resolder the shield can. Connect a 1-pF capacitor between gate 1 on the PC board and the junction of L4, C4 on the mixer board. Interconnections between boards are via RG-174/U coax. This change makes the 116-MHz oscillator common to transmit and receive

*Spectrum International, Box 87C, Topsfield, Mass. 01983.

fig. 4. Control circuits. A and B are primary control and antenna switching circuits. Circuit C switches the TR-3 exciter between receiving converter and the 10-dB pad during transmit mode. D is the attenuator necessary to reduce exciter output power.



functions. No further changes are required in the receiving converter.

A pair of 1N100 diodes connected back-to-back between receiving-antenna terminals and ground will provide front-end protection.

alignment and test

After completing the wiring shake the chassis vigorously to remove wiring debris.* Inspect all soldered connections on tube sockets, tie points, and terminals. A 5X magnifying glass is helpful here. Often an unsoldered wire will be revealed beneath top connections on tie points. A toothpick is a useful device to test the integrity of soldered connections while inspecting the work through the glass.

I follow a standard procedure after wiring circuits, which consists of cleaning all soldered connections with a 50/50 mixture of toluene and alcohol to remove solder flux and grime. Next I check all point-to-point wiring with an ohmmeter. With all tubes removed I then check voltages throughout the unit. (Voltages will be slightly on the high side with tubes removed.) These checks are worthwhile, and often a wiring error can be detected and corrected before damage occurs. A time-consuming and tedious procedure — but well worth the effort!

After completing the post-wiring checks, install the 12AT7, 6360s, and OB2s. Solder a 50-ohm noninductive 2W resistor between link output and ground. Apply power and tune the oscillator and doubler circuits for maximum output. Turn off the TR-144 and install the 4X150 in its socket.

Remove the 50-ohm resistor. Apply power and, after sufficient warmup, key the exciter. With no drive applied, adjust the bias pot for 50 mA of resting plate current on the 4X150. Apply drive and tune L10, C3 and C4 for maximum screen-current indication, then tune plate and load controls for maximum rf output. Plate current should be 200 mA. Power input with the transformer I used was 180 watts dc — power output was 108 watts as measured with a Bird Thru-line wattmeter into a Heath dummy load.

mixer output control

The TR3 tunes from 28 to 29.7 MHz. Tuning the TR144 from the low to the high end indicated a significant reduction in output power. This reduction was traced to the output circuit of the 6360 mixer. The 6360 mixer output circuit is not sufficiently broad to accommodate a 1.7-MHz bandwidth. To obtain maximum output over the entire range, I added a front-panel tuning control for C3 in the mixer output. A Rube Goldberg lashup of two flexible shafts from a panel bearing mount is shown in the photo. A machinist friend is making a miniature right-angle drive so the installation can be improved esthetically. This control allows the mixer output to be peaked over the full tuning range of the exciter.

oscillator pulling

The unit was tested on the air for about a week. A shift in frequency between transmit and receive modes was noted. Circuit analysis suggested only one possibility — the 12AT7 oscillator was pulling. Coupling between oscillator and doubler had to be reduced.

Several changes in coil positions were tried without success. An outboard oscillator using a 6CW4 was constructed, and the 12AT7 was changed to a buffer amplifier. The 6CW4 oscillator voltage is 105 Vdc regulated. Oscillator output is coupled to the 12AT7 with a 5-pF capacitor. The grid resistor in the 12AT7 was changed to 1 megohm.

The oscillator was built on a piece of scrap PC board and mounted by the retaining hardware of the driver tune-control panel bearing. When air tested again, no frequency shift was observed between transmit and receive modes. Sufficient room is available on one corner of the mixer board to mount the 6CW4 tube socket, coil, and other components. The board was bench tested and worked perfectly.

*A small bicycle air-pump is also helpful when cleaning a chassis. *editor.*

painting

Sanding with light horizontal strokes, using fine-grade emery cloth, yields a brushed aluminum finish on the panel. The shiny center strip was made by using a ¼-inch masking tape mask. The upper panel half is dove gray; the lower half equipment gray (both are Krylon colors).

Each panel half was painted with *light* coats of paint and baked for 30 minutes. Front panel lettering consists of press-on transfers — black on the light-gray and white on the dark-gray panel halves. Three coats of clear acrylic lacquer spray, baked for 30 minutes, protect the lettering.

conclusion

The TR144 completes my original project to design and build six- and two-meter transverters to match Drake TR-3 equipment. Switching the TR-3 rf connector through a Waters coax switch allows for easy selection of either six- or two-meter output.

An auxiliary rocker switch was installed on the panel to switch the TR-3 keying line to either transverter. The TR-144 faithfully transverts the TR-3 sideband quality, and on-the-air reports have been excellent.

It's quite satisfying to look over one's operating position, see identical cabinets, and be able to say, "I built it!" This is the prime source of my enjoyment in ham radio.

Once again I must express my thanks for the photographic talent and equipment of Dick McGinn, WA1IMS. I'll be happy to answer any inquiries about the construction of this unit upon receipt of a stamped, self-addressed envelope.

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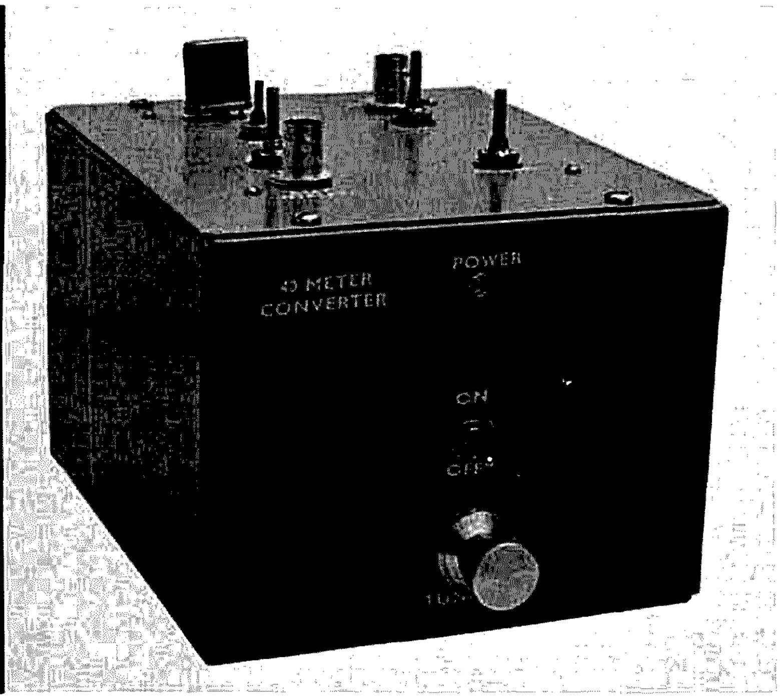
ham radio



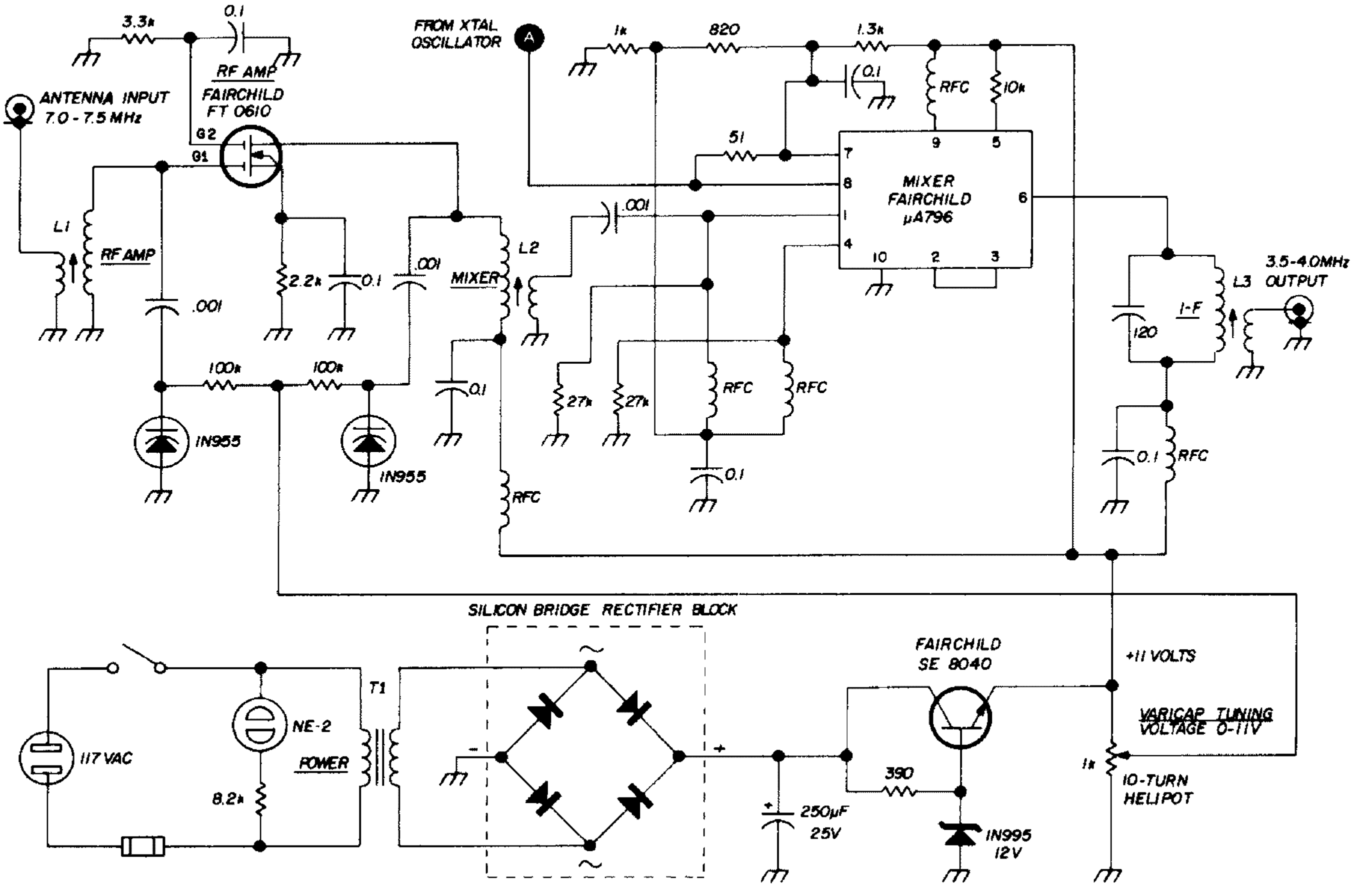
third-generation solid-state high-frequency converter

This converter combines ICs with transistors to provide optimum monoband performance

Mike Goldstein, VE3GFN

Anyone who spends a lazy afternoon thumbing through the past few years' ham publications is bound to notice the rapid advances in solid-state technology. This is true not only of commercial equipment but of the homegrown variety as well. For example, a few years ago the "in thing" was a receiver with a bipolar transistor frontend; even with the low-impedance problems and poor cross-modulation characteristics they presented. Today, to the knowing builder, they are passé; design has already run the gamut from jfet frontends through unprotected and then protected mosfet designs, and new ground is being broken using integrated circuits. How's a guy supposed to keep up to date when things move so fast?

A few years ago, I built a fet converter which seemed to work pretty well.[1] Shortly afterward, they introduced the dual-gate mosfet, which put my converter back into the Dark Ages. I went back to



40-meter circuit values

L1,L2	35 turns no. 29 closewound on Cambion 1534/2/1 form. L is 12 µH, Q is 50. The link is 6 turns of no. 27 wire on the cold end of L1.
L3	50 turns no. 29 wire closewound on Cambion 1534/2/1 form. L is 17 µH and Q is 40. The link is 6 turns of no. 27 wire on the cold end.
L4	75 turns no. 37 closewound on Cambion 1536/2/1 form. L is 30 µH and Q is 70.
RFC	1 mH at 35 mA.
T1	Hammond 166D20, primary: 120 V, secondary: 20 V centertapped at 0.1 A. Note: only one half of secondary is used.
Y1	3.5 MHz.

For 20-meter operation, parts values are:

L1,L2	15 turns no. 27 closewound on Cambion 1534/2/1 form. L is 2.5 µH and Q is 100. The link is 4 turns of no. 27 wire on the cold end.
L3	50 turns no. 29 closewound on Cambion 1534/2/1 form. L is 17 µH and Q is 40. L3 is shunted with a 5.6 k resistor. The link is 3 turns no. 27 wire on the cold end.
L4	20 turns no. 27 closewound on Cambion 1534/2/1 form. L4 has 50 pF connected across it instead of the 82 pF used in the 40-meter converter.
Y1	10.5 MHz.

fig. 1. Schematic diagram of the third-generation mobile converter for 40 meters. For 20-meter operation, parallel the primary of L3 with a 5.6k resistor and change the tuned circuits to appropriate values.

the drawing board and came up with a second generation converter.[2] Then Motorola came up with an IC double-balanced modulator (the MC1596G) which turned out to be a superb mixer, and I was done in again![3]

This third-generation converter, unlike previous efforts, is a single-band converter

designed to deliver optimum performance on 40 meters for net operation during the poor summer conditions. Low noise, high gain and good dynamic range were all important design requirements, as well as the usual high-stability and image-rejection performance necessary for smooth operation. The circuit can be used to cover any band right up to ten meters merely by substituting the proper tuned circuits and crystals; a version of the same circuit has been used at 100 MHz with no noticeable instability. Making such a converter bandswitching would be quite easy.

circuit

Varactor diode tuning is used in the rf amplifier which uses a Fairchild FT0601 dual-gate protected mosfet. The RCA 40673 could be used as a direct substitute. The tuning is very smooth, with no instability of any sort, even though the amplifier is not neutralized. It is possible to peak the converter on a signal anywhere in the band, but the tuned circuits are broad enough that constant retuning with frequency change is unnecessary. This is a refreshing change from previous efforts in which all circuits were deliberately resistance-loaded to broadband them, which compromised the stage gain and increased the chance of images appearing.

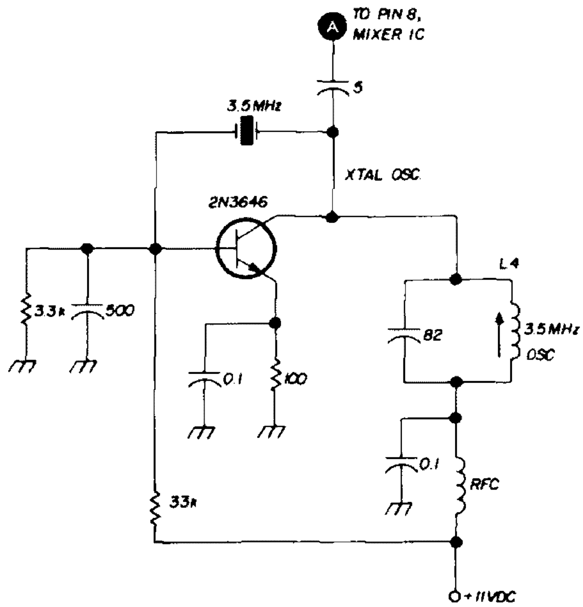


fig. 2. Schematic of the local oscillator for the hf converter. For 20-meter operation change the 5 pF oscillator-output coupling capacitor to 2 pF and change the crystal to 10.5 MHz along with re-resonating the tuned circuit.

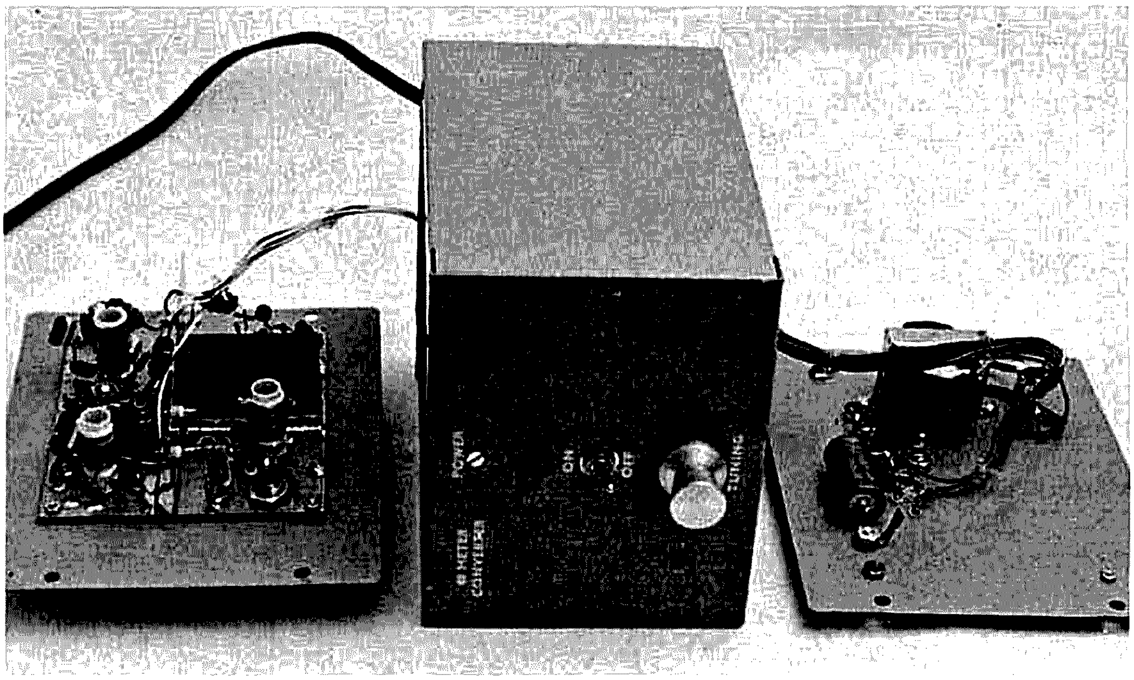
"Each kit has a vital component or two missing."

A double-balanced modulator was used for the mixer. The Fairchild μa796, or the Motorola MC1596G or 1496G may all be used in the same circuit. These IC devices make excellent mixers, with *only* the sum and difference of the two input frequencies appearing at the output. A lot of the spurious rejection takes place in the mixer which seems to actually refuse to process anything other than the correct input coming from the rf amplifier. The chance of any spurious crossproducts appearing in the output is very low. In fact, the output waveform is such a clean sinusoid (at 3.5 to 4.0 MHz) that it can be used to drive a digital frequency counter – an instrument easily confused by input waveforms containing more than one frequency.

The oscillator is an old reliable circuit, and almost any npn high-frquency transistor can be used. I used the 2N3646 for this unit, but the 2N706, 2N4124 or

HEP50 would work as well, and the circuit should work right up to ten meters even with overtone crystals.

You can use any low-power npn transistor in the series-pass regulator in the power supply. The design is simple, and you can use any bridge block rectifier or discrete diodes. The converter draws only 7.8 mA from a 12-volt supply, so the unit could be powered by a 9-volt transistor-radio battery for quite a while without battery deterioration. If battery operation is likely, use a 100k tuning potentiometer.



Overall view of the 40-meter converter board, cabinet, power supply and connecting wiring.

construction

The converter is built directly on a piece of glass-epoxy printed-circuit board using press-fit terminals where needed and with shields and ground connections directly soldered to the copper. This type of construction is excellent for rf work, eliminating the usual mess of nuts, screws and lugs needed for grounding to an aluminum chassis, and the time and materials needed for printed circuitry. An alternative would be to use the new *Circuit-Stik* instant printed-circuit decals. None were available at the time of building. The type of construction used added to circuit stability, allowing very short leads and solid ground connections. Nothing less than a good hot soldering iron should be used to make all ground connections, as a cold joint anywhere could be fatal to circuit stability.

All semiconductors are soldered directly into the circuit, using a heat sink when soldering each lead. Keeping leads short and eliminating the possibility of lead-to-socket capacitance assures stability without neutralizing — even on ten meters.

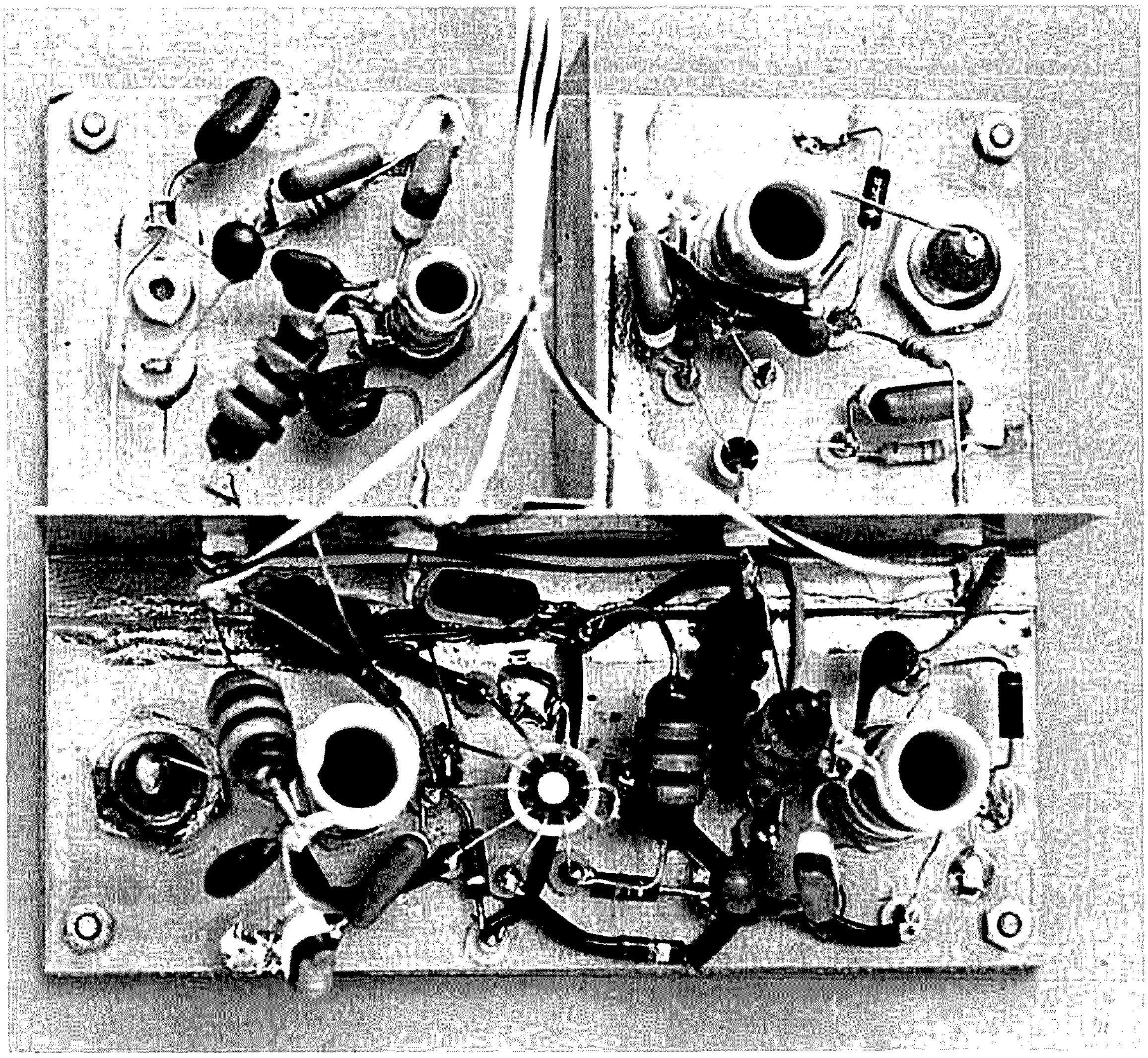
The general layout of the circuit board can be easily seen in the photograph. The layout minimizes stray capacitance and isolates inputs from outputs; it should be followed quite closely.

To make sure the holes in the circuit board line up with the utility-box cover upon which the board is mounted, the board was mounted on the cover and all centering holes for coils and connectors were drilled through both; the circuit board was then removed from the cover while the circuitry was installed. In this way the holes for the terminals are

concealed when the final assembly is completed.

When construction is completed, set the varicap tuning voltage at the potentiometer wiper at 7 volts, tune in a signal at 7.15 MHz, and adjust the rf amplifier, mixer, and i-f coils for maximum signal. You're in business.

The converter input and output impedance, with the coils constructed as shown, is 50 ohms. When driven and terminated in this impedance the system gain is about 30 dB, so keep the receiver gain low to avoid overload. If the converter has too much gain for the receiver used, insert a resistance (1000 ohms or more) between pins 2 and 3 of the mixer IC. Note that the 3.5-MHz crystal oscillator will be heard when the receiver is tuned to 7 MHz.



The 40 meter converter circuit board. Upper right section: rf amplifier. The Varicap can be easily seen just to the right of the antenna coil. The FT0601 is in the lower left area of the section, upside down, with the drain connected directly to the feedthrough terminal. Upper left section: oscillator. Lower section: mixer. The IC is turned upside down and mounted directly by its terminals. Pins 2 and 3 are clipped short and soldered directly together. Note how all ground connections are soldered directly to the copper board.

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ham radio

pre-emphasis for ssb transmitters

Reducing peak-to-average power ratio results in increased talk power — here's an af processor that gives 6-9 dB improvement in ssb rigs

V. Aumala, OH2CD, Vanhaistentie 4B7, Helsinki 42, Finland

Single-sideband transmitters are usually rated in terms of peak envelope power (PEP). According to one authority,[1] the average output power of an ssb transmitter is one-half the peak envelope power if no distortion is present. This statement is made on the basis that measurements are taken when the ssb transmitter is modulated with a standard test signal of two equal-amplitude signals.

It can be shown that the ordinary voice contains high-amplitude peaks that are about 14 dB greater than the average level.[2] Furthermore, it is well known that ssb transmitters are peak-power limited. This means that an ssb linear amplifier may be able to handle 1000 watts average power but will begin to flat top when peak power approaches 2 kW.

From the considerations above, it may be inferred that many amateur ssb transmitters are not being used to their full capability. If the peak-to-average ratio of the ssb transmitter output power can be reduced, then the talk power, or articulation index, can be increased. Speech processing circuits have been introduced to reduce the ratio of peak-to-average power; however if the processing scheme

introduces distortion, then the articulation index is reduced. The best method of speech processing, for ssb transmitters, is one that provides maximum increase in talk power without distortion and subsequent flat topping.

Let's look at two popular methods of speech processing in amateur ssb equipment: the speech compressor and speech clipper.

compressor

The speech compressor, which is used in the audio circuits of the ssb transmitter, has a typical attack/release time constant of 5×10^{-3} and 5×10^{-1} second. As the ratio of these time constants is made shorter, the compressor action approaches that of a clipper. However, audio compressors do not affect the peak-to-average power ratio of the ssb transmitter; therefore we'll not discuss them further.

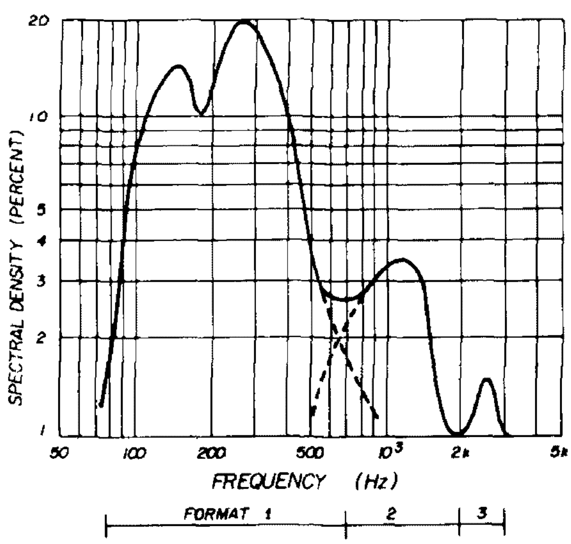

fig. 1. Profile of speech spectral density.

clipper

A speech clipper affects the ratio of peak-to-average power in an ssb transmitter, which is of interest here. First, let's see what is clipped. Speech spectral density is distributed according to **fig. 1.** This curve isn't quite accurate, because speech characteristics vary with differences in languages and with male and

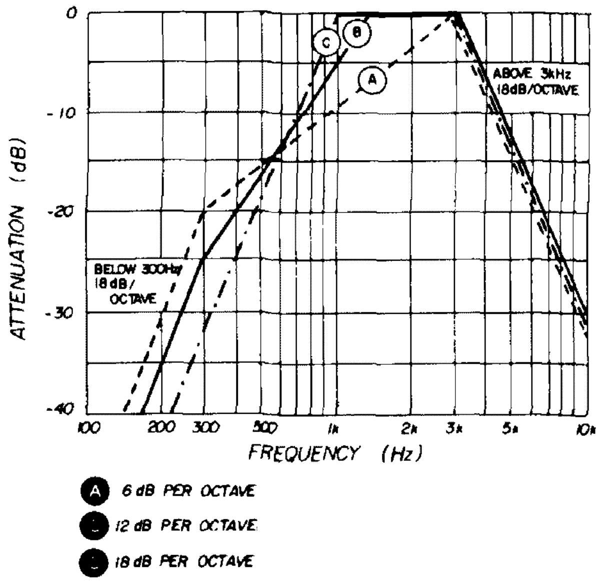

fig. 2. Results of different pre-emphasis networks analyzed in reference 3. Optimum performance was obtained with curve B.

female voices; however it is sufficiently accurate for our purposes.

Note that spectral density is distributed into three formants. The first formant is dominant and lowest in frequency. This formant includes most of the sounds in personal voice characteristics. Such voice power has little to do with readability in an ssb radio circuit; it merely adds fidelity to the transmission. Audio-frequency compressors emphasize this speech-energy formant, which is why they don't do much good in ssb communications systems. When clipped, the energy envelope in formant 1 creates strong harmonics, which create harmonic distortion.

Formant 2 contains most of the communications intelligence, but it is of much lower power level. Reference 3

examines different types of pre-emphasis circuits and gives optimum slopes of audio-response curves. From fig. 2 it is seen that the lower in frequency at which tioning by a low-pass filter, noise was added until the signal-to-noise ratio was unity. The criterion for evaluating test results was the percentage of words

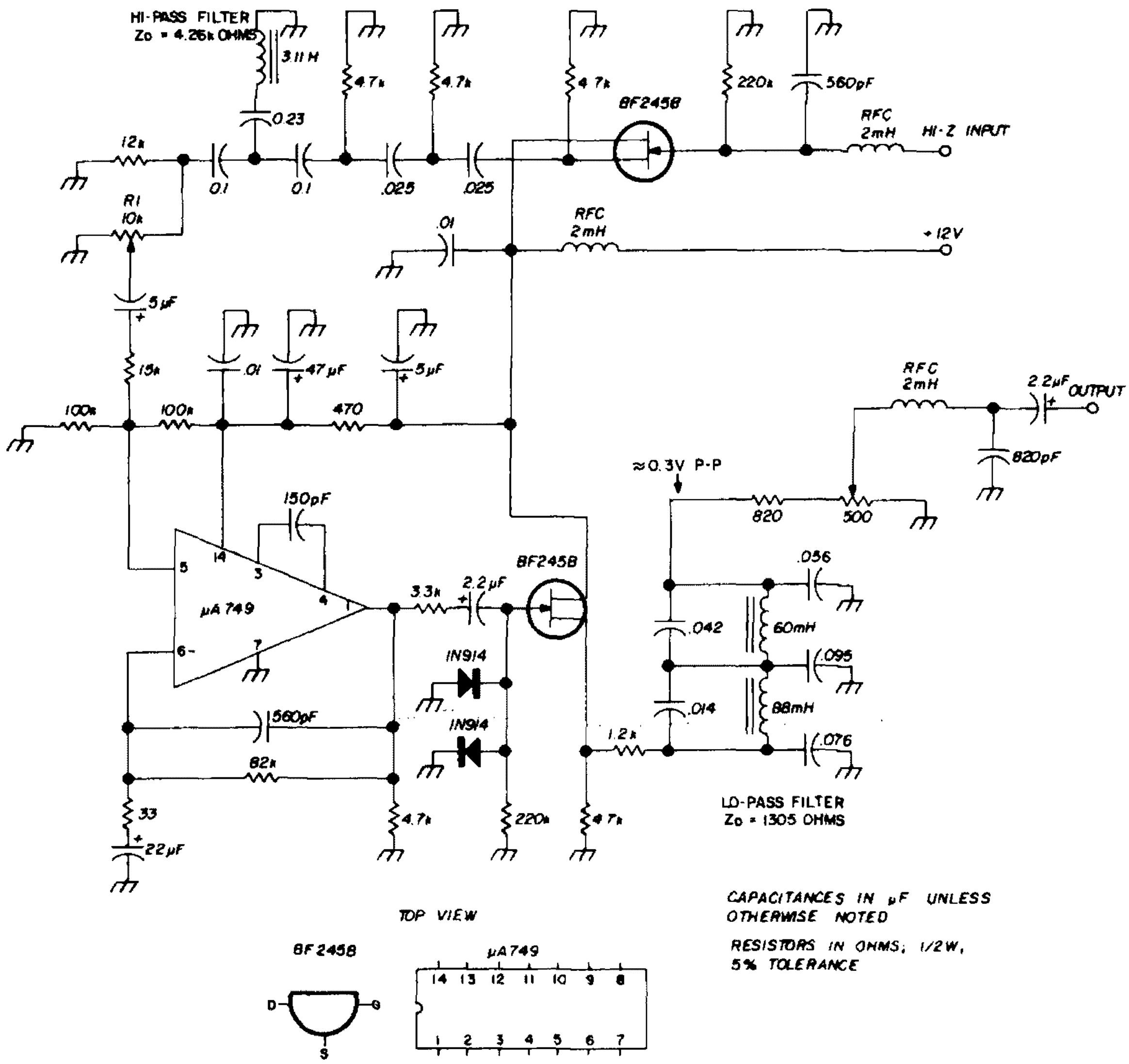

fig. 3. Pre-emphasis circuit used by the author. High-and low-pass filters are those shown in reference 4. Diodes should be matched for symmetrical clipping action. Inductors in the low-pass filter are toroids, available from many surplus sources. Capacitor values are important for proper operation; correct capacitances may be obtained by combining standard units. Pot R1 adjusts clipping level.

the knee of the response curve occurs, the steeper will be the slope. Best results have been obtained with curve B of fig. 2, where the kneepoint is at 1.1 kHz. The slope of this curve is 12-dB/octave.

test results

Tests were made as follows. Pre-emphasis was first applied to the signal, then the signal was clipped with a Schmitt trigger. After clipping and conditioning by a low-pass filter, noise was added until the signal-to-noise ratio was unity. The criterion for evaluating test results was the percentage of words understood in the transmission. Test results were:

1. Linear nonclipped speech – 10 percent.

2. Linear response with "infinite" clipping – 60 percent.

3. Pre-emphasis and clipping – 90 percent.

The most interesting observation was

that no equalizer was needed after the processor (e. g., in the receiver).

The benefits of pre-emphasis circuits are most noticeable in af clipping. Pre-emphasis circuits reduce the harmonic content in the microphone channel, whereas rf clippers must use elaborate filters to accomplish the same objective at radio frequencies.

pre-emphasis circuit

The schematic is shown in fig. 3. The input circuit uses a source follower for high-impedance microphones. If you have a low-impedance microphone, this stage may be omitted. Two RC networks in the input circuit shape the slope for a response of 12 dB/octave below 1.4 kHz (see fig. 4).

A high-pass filter[4] attenuates frequencies below 300 Hz. A low-noise, high-gain amplifier (Fairchild μA749) increases the microphone signal for the clipping diodes, a pair of 1N914s. The degree of clipping can be controlled with a pot at the amplifier input (R1, fig. 3).

The diodes must be matched pairs to obtain symmetrical clipping action.

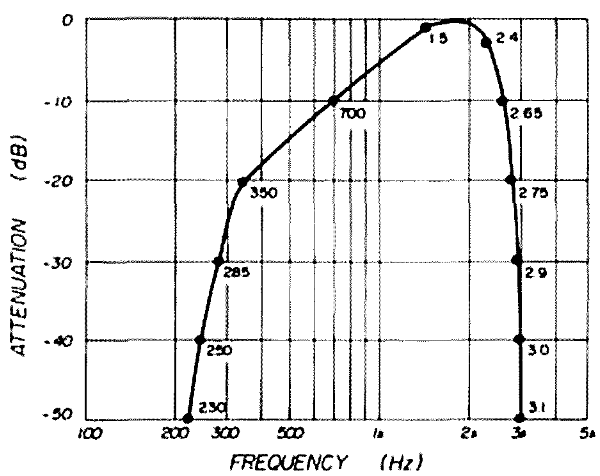

fig. 4. Speech-processor response without clipping. An optimum slope of 12 dB/octave below 1.4 kHz is obtained with two RC networks in the imput circuit.

A source follower is used after the clipper to obtain the proper impedance for a low-pass filter. This filter, also shown in reference 4, reduces frequency response above 2.4 kHz and attenuates harmonics. The processor output is sufficient for most solid-state balanced modulators.

conclusion

I have used this processor in a solid-state 10-watt mobile transceiver with

The speech processing unit described in this article is included in the 10 watt transceiver shown in this photograph of the author, OH2CD.

encouraging results. Approximately 20 dB of clipping resulted in 6-9 dB output improvement in speech intelligibility.

Thanks are due OH1OY, who inspired me to build this unit while a larger unit, using rf clipping, was being developed. But that's another story, as Kipling used to say.

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ham radio

modular receiver

for two-meter fm

Modular construction with inexpensive kits produces an easy-to-build two-meter fm receiver or converter

Harvey L. Wagner, WA2GBF, 83-19 141 Street, Jamaica, New York 11435

Over the past few years vhf fm has moved from the private domain of a few hardy experimenters to an increasingly popular mode of communications. One drawback still exists for amateurs who wish to get on fm: good equipment is either expensive, or as in the case of surplus gear, takes modifications which are enough to discourage the average appliance operator from digging in and learning what radio is all about. Because of these factors – expensive equipment on one hand and apparent complexity on the other – I set out to find a reception method that would give the average amateur a chance to hear what fm is really like before making the commitment in time and money that current equipment demands.

receiver design

Keeping in mind that the receiver must be simple enough for an appliance operator to build and inexpensive enough so that even an out-of-work college student (like myself) could afford it, modular construction was chosen as the simplest and cheapest system on which to base a workable receiver. The heart of this receiver (or hearts, I should say) are transistor module kits from International Crystal which tremendously simplify construction and subsequent modifications of the unit. The design uses an fm broadcast receiver for i-f, detector and audio output stages; leaving only the

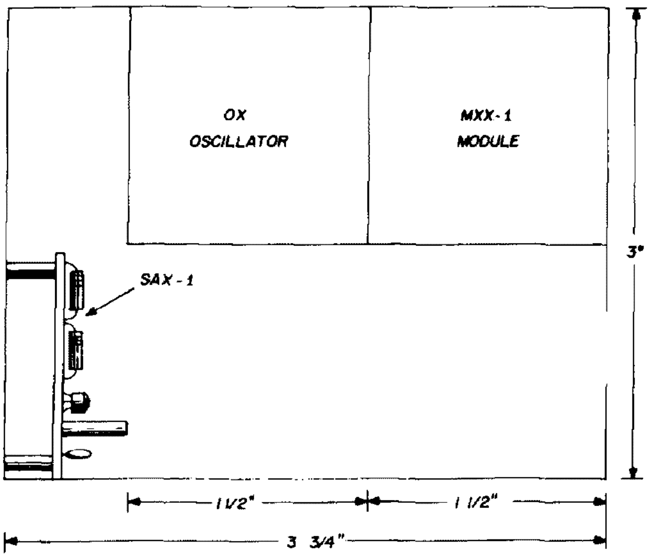


fig. 2. Converter is built into an ordinary minibox.

front end, mixer and local oscillator stages for the builder. If you wish, this system can be built as a converter to be connected to an existing broadcast receiver or it can be built as a complete receiver utilizing the circuit board from the now ubiquitous Japanese a-m/f-m radio.

Fig. 1 shows the block diagram of the basic modular receiver. The SAX-1 module which is used as the front end, is a small-signal amplifier with a sensitivity of 1 microvolt and about 10 dB gain at two meters. The MXX-1 module, designed specifically as a mixer, has a conversion gain of about 6 dB and a sensitivity of one microvolt. The local oscillator in this circuit is the OX oscillator module with a type EX crystal.

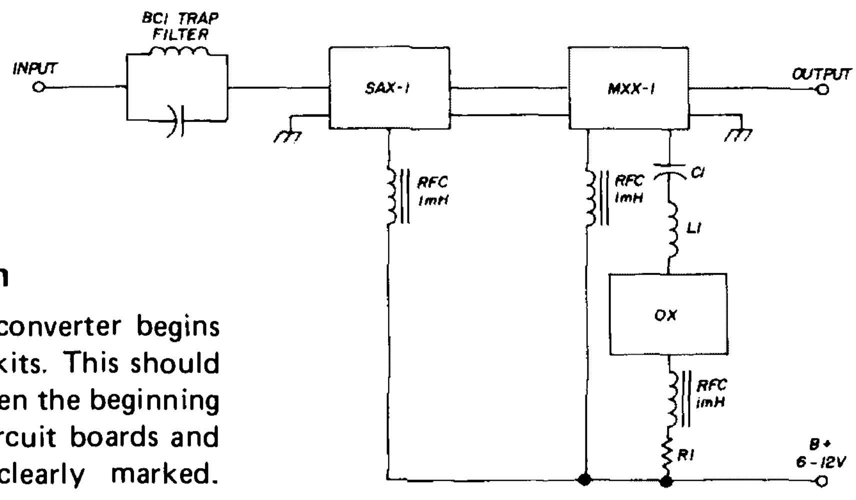


fig. 1. Block diagram of the complete converter.

converter construction

Construction of the converter begins with wiring the module kits. This should present no problem to even the beginning builder as they are all circuit boards and have parts placement clearly marked. When wiring the circuit boards, the builder will have to select the proper coil and capacitor combination for the tuned circuits from several supplied. When making this selection, bear in mind the input and output frequencies of the converter and the crystal frequency needed to produce this conversion.

Assembly of the converter section is simple and straightforward. The circuit is housed in a **2** x **3** x **3¾** inch minibox, and suggested placement of the modules within the box is illustrated in **fig. 2.** The SAX-1 module is mounted vertically at one end of the box next to the input jack. The OX and MXX-1 modules are mounted horizontally with the MXX-1 closest to the output jack.

Connections carrying rf between modules should be made with small shielded cable such as RG 174/U, and all power leads should be choked to prevent the circuit from oscillating. The value of resistor R1 in the B+ lead to the OX module is determined by the voltage used to power the converter and can be found in **table 1.**

The selection of a crystal frequency for the converter will depend on the output frequency used. For the 88- to 108-MHz band, the frequency of the crystal may be found by using the formula $f_{crystal} = f_{in} - f_{out}$. For output frequencies below 88 MHz the crystal frequency is found by using the formula $f_{crystal} = (f_{in} - f_{out})/2$. If outputs below 88 MHz are used, it will be necessary to use the series tuned circuit L1 – C1

table 1. Supply voltage vs R1.

voltage	R1 (ohms)
6	0
9	150
12	300

which is tuned to the local-oscillator injection frequency, which is the second harmonic of the crystal frequency. The values of L1 and C1 will depend on the injection frequency, and can be determined by the use of a nomograph such as the one which appears in all recent issues of the *ARRL Handbook*. For injection of the fundamental crystal frequency, L1 and C1 may be omitted, using shielded cable between the OX and MXX-1 modules.

In some areas, it may be necessary to install a parallel-tuned trap filter at the antenna input of the converter to eliminate feedthrough of local fm broadcast stations. The filter should be tuned to the frequency of the offending station.

Before power is applied, *all* tuned circuits should be adjusted to their proper frequencies with a grid-dip meter. Although this procedure is not absolutely necessary, it is a good idea as it makes alignment much simpler, and if an error has been made in component values in a tuned circuit, this procedure will allow you to *detect and correct it* before things reach the hair-pulling stage. If you do not own a grid-dip meter it may be a good idea to buy or build one as it is one of the most useful pieces of test equipment you can have around the shack.

If the converter is to be used as part of a receiver rather than as an outboard converter it is worth pointing out that most, if not all, of the imported transistor radios use a positive ground system; this converter uses a negative ground. If the receiver board has a positive ground, care should be taken to isolate the board from the chassis ground used for the converter. It may be simpler to power the receiver board with a separate battery and to leave it in its plastic case.

operation

When power is applied to the converter there will be a slight increase in receiver background noise. If there is an extremely high noise level and all manner of whistles, blurps and birdies appear, or if the converter draws more than about 35 mA, the circuit is probably oscillating. If oscillation occurs and the power leads have all been choked, it will be necessary to shield the MXX-1 module with grounded partitions made of flashing copper or copper-clad circuit board.

When the converter is operating properly it is aligned as follows:

1. Using a signal generator, grid-dip oscillator or received signal, peak the coils on the SAX-1 and MXX-1 modules for maximum audio output at the receiver.

2. If a tuned circuit is used between the OX and MXX-1 modules adjust it for maximum audio output at the receiver.

3. If you used a trap filter tune it for maximum attenuation of the interfering signal.

4. All of these adjustments are slightly interlocking, and it may be necessary to repeat this procedure several times until maximum performance is obtained from the converter.

The prototype receiver, which consisted of the converter feeding a hastily repaired GE portable receiver, performed better than expected. Although the receiver has a wide i-f and uses a wideband detector, the only difficulty encountered in copying narrow-band signals was that of adjacent-channel interference from strong signals during crowded-band conditions.

This design is an adequate receiver for monitoring the local repeater. For higher performance several modifications can be made to the original design. The addition of an fet preamp will provide more sensitivity and greater overall gain. It may also be desirable to include a second SAX-1 module between the MXX-1

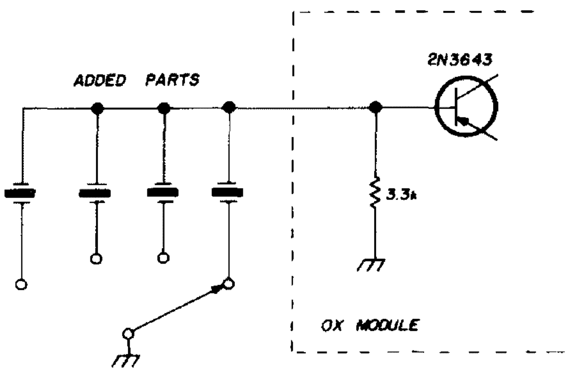


fig. 3. Fixed-frequency modification.

module and the receiver if the converter does not provide sufficient gain.

You could add a noise-operated squelch and an improved tuning mechanism. Although the receiver can be tuned with its main tuning capacitor, the designed bandspread of 20 MHz does not lend itself to tuning easily between channels spaced 60 kHz apart. You can add a small variable capacitor in parallel with the main tuning capacitor to provide a slower tuning rate over a limited range of frequencies. Another method which may be used is shown in **fig. 3**. This method uses switch selected crystals to determine frequency and guarantees that the receiver will always be right on frequency and will allow the receiver to find and monitor an inactive channel. Although I did not try either of the above methods, there is no reason why they should not work as long as you use good construction techniques and keep lead lengths short.

Transistor modules provide an easy way to get on fm. The flexibility provided by this method of construction encourages experimentation and optimization of designs. The receiver design presented in this article is only one of many possible ways for the amateur to get decent performance at two meters for reasonable cost. It is hoped that the design presented here will not be considered a cut and dried cookbook recipe, but rather a starting point for experimentation with various receiving techniques.

ham radio

a simple crystal checker

An old circuit with new components makes a handy one-evening project for testing surplus crystals

Hank Olson, W6GXN, Box 339 Menlo Park, California 94025

Often I have wanted to check a crystal for activity and frequency before using it in an electronic project. Many surplus crystals have two or three numbers on them, any one of which could be the crystal frequency; and it is sometimes unknown *which* mark to believe. By plugging the unknown crystal into the checker described here, it will oscillate at its first mode, allowing me to tune it in on a nearby hf receiver. Also, a high impedance voltmeter connected to the test point of the checker gives a dc indication of oscillation.

The circuit of the crystal checker is presented in **fig. 1.** It is the old familiar Pierce Oscillator, with an insulated-gate fet substituted for the original vacuum tube. Unlike a vacuum tube, however, an insulated-gate fet cannot draw dc grid current, and so D1 has to be added if we intend for the grid-leak bias system to work. (Perhaps "gate-leak" bias system is more correct in this case.) The diode should be fast enough to rectify up to 20 MHz, I suggest three devices in order of

their preference. The Hewlett-Packard HP 5082-2800 and the Motorola MBD 501 are both Schottky Barrier (or hot carrier) types, and the Fairchild FD 700 is an extremely fast silicon switching diode.

The MFE 3004 insulated-gate fet is best *soldered* into the circuit so that it cannot be destroyed by static voltages between the gate and either source or drain. The MFE 3004 comes with a small metal sleeve that shorts all four of its leads together. Before removing this sleeve, the circuit should be finished and ready to receive the fet. Then wrap several turns of fine tinned wire (one strand pulled from a short piece of stranded hookup wire) around the MFE 3004 leads, so as to short them all together. Remove the metal sleeve and solder the fet into the circuit, leaving the shorting wire in place until finished soldering.

The Pierce circuit is fundamentally a form of the Colpitts oscillator, where the crystal looks inductive, and parasitic capacitances of the fet form the capacitive tap. This is shown in fig. 2; note that

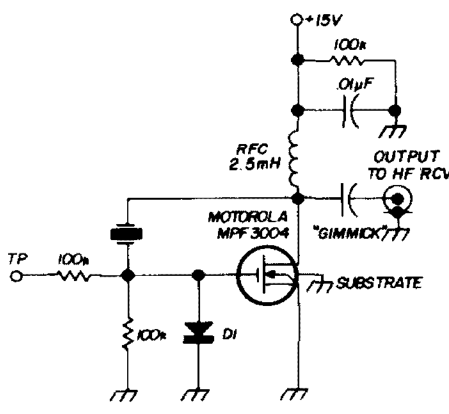


fig. 1. Schematic of the crystal checker. D1 is explained in the text. The checker is made for fundamental mode operation only.

C_{gd} is in parallel with the crystal, and the net reactance of the crystal and C_{gd} is inductive.

The crystal checker will only work with fundamental mode crystals in the 1- to 20-MHz range. Below 2 MHz, the impedance of the rf choke becomes too low to allow the drain to appear to be floating. AT-cut crystals are not generally made for the fundamental mode above about 20 MHz. Many surplus crystals in

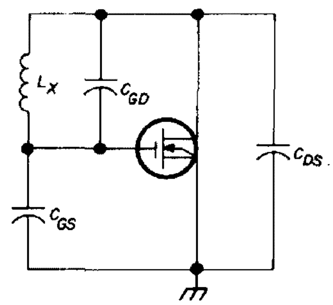


fig. 2. Equivalent circuit of a Pierce crystal oscillator, showing that it is, in fact, a form of Colpitts oscillator.

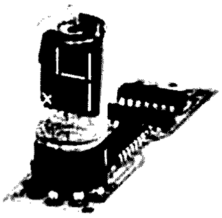
the 20-MHz range (particularly CR9/U and CR24/U) are third-overtone types. If such a crystal is used with the checker (with appropriate pinning adapter) it *will not* oscillate as marked, but at approximately one-third that frequency. For instance, a CR9/U crystal marked 19.825 MHz oscillated at 6.615 MHz.

As shown, the crystal checker was built to accommodate either the FT 243 or HC6/U styles of crystals, because these are by far the most common. Other crystal sockets could, of course, be added. The checker is constructed in an LMB-00 box chassis with the battery terminals out one side. Since I only use the unit occasionally, no power switch was included; I simply connect the 15 volts for use. A Burgess U10 or Eveready 411 battery is adequate as a 15 volt source.

With the 15-volt battery connected, the fet will draw I_{dss}, which is rated 2 to 10 mA for the MFE 3004. A high-impedance voltmeter (such as a vtvm) measuring the voltage between ground and test point should read about zero before a crystal is plugged in. When the crystal is plugged in, the voltage at the test point should jump to some negative voltage (−10 volts for a good, active crystal) indicating oscillation.

ham radio

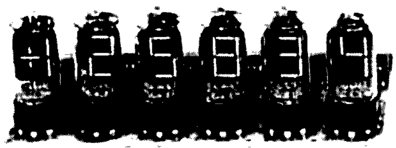
epd ENVIRONMENTAL PRODUCTS



UNIVERSAL DCU/DRIVER

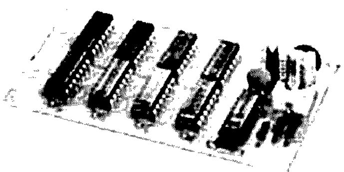
These universal counting modules have heavy duty outputs to drive all 7 segment displays requiring up to 15V and 40mA per segment. TTL used throughout. Requires 5 volt @ 120mA per module and any number of modules may be cascaded. Typ. count rate is 20MHz except the NR-3H which is 70 MHz. Will drive any display in this ad (not included).

NR-3	Modulo 10 Counter 20 MHz	$8.75
NR-3A	Modulo 6 Counter 20 MHz	8.75
NR-3H	Modulo 10 Counter 70 MHz	10.50
NR-3B	Modulo 12 Counter clock	12.95



The most versatile multi-digit counting kit ever offered!! Comes with 5½ digits of counting and 7 segment decoding with the same heavy duty outputs as the NR-3. Typ. maximum count rate is 70 MHz. Each counting stage is completely independent allowing a custom approach to your application. The "+" and "—" sign are great for DVM or over-under circuitry. The overrange stage also has an overflow latch and reset buffer. Displays not included. PC board is 3½" x 6".

NR-3FM Multi-Stage Counting Unit $39.95



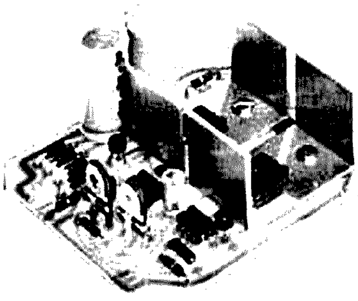
CRYSTAL FREQUENCY STANDARD

Contains a .002% crystal oscillator with TTL decade dividers to give output frequencies of 10, 1 MHz, 100, 10, 1 kHz, 100, 10, 1, & 0.1 Hz. Kit requires 5 volt supply @ 175mA. Uses low TC components and has zero-beat trimmer. Great for freq meter, digital clock, etc. W/ complete instructions. CRO-1P $21.95

LINE CLOCK SOURCE

The DCC-2 derives precision gating and clock signals from the 60 Hz line. The input is a combination schmidt trigger and integrator which eliminates false triggering from line noise. The input is over-voltage protected and requires no adjustment. TTL compatible output frequencies are 10, 1, 0.1, & 0.01667 (1 pulse/min.) Hz. PC board measures 1.2" x 3.5".

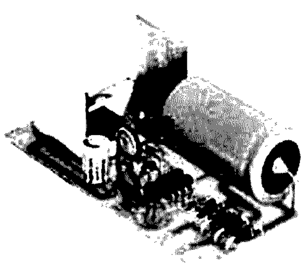
DCC-2A Line Frequency Standard $12.95



ANALOG POWER SUPPLY

Power your OP-AMPS with this versatile, low cost, dual-polarity regulator. One control varies both outputs simultaneously over the range of 0-15V. Electronic current limiting may be set separately for each output over the range of 25-200mA. Regulation is 0.1% and the ripple is below 3mV RMS. All parts conservatively rated for long life.

APS-5A Op-amp Power Regulator	$12.95
TR-200 Transformer for APS-5A	2.95

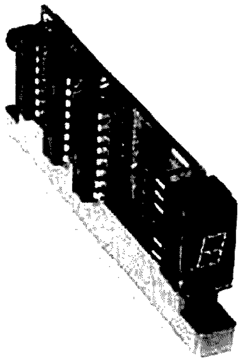


LOGIC SUPPLY REGULATOR

Both kits have an output range of 3.3V to 5V with current limiting and short circuit shut down. Regulation is 1% and ripple & noise is 10mV RMS. Heavy duty components insure long life and allow rugged use.

DPS-1A Output current 0.6A	$6.95
DPS-2A Output current 2.2A	9.95
TR100 Transformer for DPS-1A	2.29
TR500 Transformer for DPS-2A	5.95

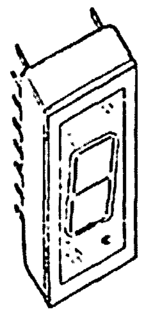
More Details? CHECK—OFF Page 94



LED READOUT MODULE · COUNTER · LATCH · DRIVER

This new module has all the most desirable features required in counting and display. Fits 18 pin connector. 5V supply. All TTL logic. 20 MHz count rate.

$16.95



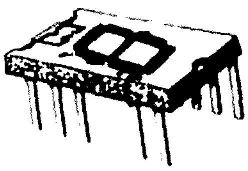
Miniature 7 segment display mounts in 16 pin dual in-line socket. 5V operation at 8 mA. per segment. 100,000 hr. life. W/decimal pt.

3015 Miniature Display .... $3.45 3/$10.00

LARGER 7 segment display as pictured with the NR-3 series kits. Bright numerals can be seen even in direct sunlight. Mounts in 9 pin miniature socket supplied with the kits.

2010 7 Segment Display 0-9 DP $4.45

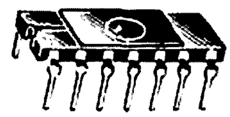
2020 As above except "1" "+" "—" 3.95



LED

Electrically and physically identical to the popular MAN-1. Has higher light output than MAN-1. Includes LH decimal point. W/spec sheet & app information.

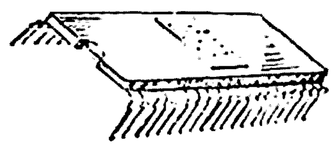
LED700 7 Segment LED Display $7.95
(QUANTITY DISCOUNTS AVAILABLE)



RAM

The 1101 Random Access Memory (RAM) will store and readout 256 bits. The chip is TTL compatible and comes with a complete spec sheet w/applications.

1101 256 Bit RAM $8.95



LSI

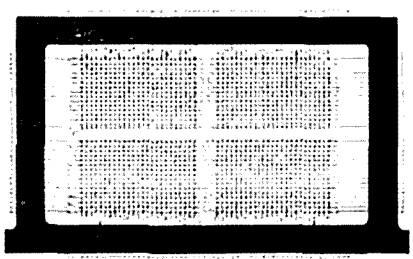
Build several instruments with this chip and little else. First really useful LSI chip for the experimenter. Contains:

- Four decade counters + overrange
- Four 4 bit latches w/BCD outputs
- Seven segment decoder
- Display multiplexing circuitry
- Two programmable *oscillators*
- Single 5 volt supply at 5mA ! ! !
- Inputs TTL compatible
- Housed in 28 pin dual in-line pak

Comes with 9 p. booklet of specs & app notes. Booklet contains interfacing info on all seven segment displays made.

5002 LSI Chip $26.95

5005 Booklet on 5002 only 1.50



At last! Noncritical memory planes for the experimenter. Made by Ampex for IBM spares. They were removed from NEW core stacks. The large 50 mil cores allow the use of the most inexpensive sense amps. The cores are in an 80x50 array. All the necessary core specs are included with each plane. Available is an 80 page booklet describing an 8 bit x 1000 word memory using the MP-2A. Parts lists, schematics, and app notes are included in the booklet.

MP-1A 4000 Bit Core Plane $12.95

MP-2A 16K Bit Core Stack 47.50

MPB-1 80 Page Core Memory Booklet 10.00

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More Details? CHECK—OFF Page 94

calculating the inductance of toroids

Two handy formulas and a table give quick approximations of the inductance of homemade toroids

Michael J. Gordon Jr., WB9FHC, 203 Woodbine Avenue, Wilmette, Illinois 60091

Toroid inductors, with their high Q, small size and self-shielding properties, are excellent for use in modern solid-state gear – especially where space is a valuable commodity. In the past, however, there has been one drawback: It was very difficult to calculate how many turns it would take to give you a desired inductance. In a back issue of *QST*, there were instructions for doing this, provided you owned an *ARRL Lightning Calculator Type A*, or a *Coil Winding and L/C/F Calculator Type A.*[1] I did not have one of those devices, so, rather than waiting until I could get one, I sought a formula which could be used without any special apparatus.

I found the equation:

$$N = K\sqrt{L} \qquad (1)$$

Where N is the number of turns, L is the desired inductance in microhenries and K is a constant dependent on the toroid being used. **Table 1** lists this constant for a number of Amidon cores in common use. In this table, K is printed to eight decimal places. It is not necessary to use all of these places, and your results will be close even if you truncate the last 6 or 7 digits. For example, suppose you need a 10-μH coil and have a T-68-2 core. Substituting the known values into the equation, you obtain

$$N = 13.71\sqrt{10} = 13.71(3.162) = 43.35102$$

Since there is no such thing as a fraction of a turn on a toroid, you should round the answer off and use 43 turns.

If you desire to use a core other than one of those in table 1, you can calculate K for yourself. The formula is

$$K = \frac{N1}{\sqrt{L1}} \qquad (2)$$

Where N1 is the number of turns on the particular form you are using which gives the inductance L1. To illustrate, suppose

table 2. Maximum number of turns of various gauge wires on standard *Amidon toroid* cores.

T-94	T-80	T-68	T-50	T-37	T-25	T-12	T-05	wire size
15	14	10	8	5	3	—	—	10
20	18	13	10	7	3	—	—	12
25	22	16	13	9	4	—	—	14
32	28	20	17	11	6	—	—	16
41	36	26	21	14	8	3	—	18
51	45	33	26	18	10	4	—	20
64	57	42	33	23	13	6	—	22
80	72	53	42	29	16	8	3	24
101	90	66	53	37	20	10	4	26
127	113	83	67	47	26	13	5	28
158	141	104	84	59	33	16	7	30
198	176	130	105	73	41	20	8	32
250	223	165	133	93	53	26	11	34
307	273	202	163	114	65	32	13	36
393	350	259	210	147	83	41	18	38
495	441	326	264	185	105	52	23	40

you have a toroid form of unknown properties and you need a specific inductance. The first thing to do is put on a test winding and measure it's inductance.

The easiest way to do this is by making a tuned circuit with the test coil and a capacitor and finding the resonant frequency with a grid-dip meter. I have experimented with this and have found that the best way to couple to a toroid is by using rather long leads on the capacitor and draping the tuned circuit over the gdo pickup coil. An alternate method is to make the tuned circuit with the shortest leads possible and putting a length of wire through the core and shorting the ends. Coupling to the tuned circuit is made through this pickup loop.

table 1. Toroid-core constants for use in equation 2.

type	K
T-94-2	10.87375857
T-80-2	13.09481019
T-68-2	13.71166566
T-50-2	13.49065790
T-37-2	15.09667411
T-25-2	16.76244696
T-12-2	21.17423645
T-94-6	11.64825226
T-80-6	14.54857791
T-68-6	14.61045410
T-50-6	15.31238723
T-37-6	17.48997890
T-5-6	18.97143316
T-12-6	23.75741463
T-50-10	16.71056534
T-37-10	19.07988330
T-25-10	20.87788877
T-12-10	27.84454642

In any case, once the resonant frequency is known, the inductance of the trial winding can be calculated. Now you are ready to use **eq. 2.** Simply substitute in the now known values of N1 and L1. In case you need it, the formula for inductance is

$$L = \frac{25330}{F^2 C}$$

Where F is in MHz, C is in pF and L is in μH.

One other thing you might need to know is how many turns of a given size wire you can fit on a given toroid. **Table 2** gives this information for most of the toroids in **table 1.** In general, use the largest size wire you can, since this will help to insure maximum Q. At first glance, it might seem that many of the cores on **table 1** are not listed in **table 2.** Keep in mind that the first two numbers in the type designation specify the size of the core (a T-94-2 has the same physical dimensions as a T-94-6 and will hold the same amount of wire).

These formulas should provide the correct answers to the nearest turn or two — accurate enough for most amateur work.

reference

1. "Technical Correspondence," *QST,* April, 1971, page 48.

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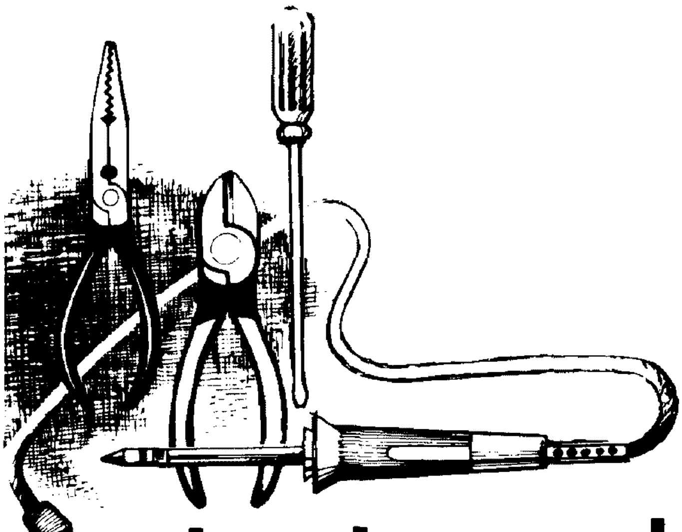


circuits and techniques

ed noll, W3FQJ

fm deviation measurements

In the enjoyable task of learning technical techniques, the knowledge gained from making tests and measurements is especially rewarding. This is particularly true when learning the makeup of the fm wave. In spite of this, very few amateurs make deviation measurements, and the carrier-zero technique of deviation measurement has been almost ignored by most radio amateurs. It should be better known.

A complete fm wave has a constant amplitude and a varying frequency. It is composed of a carrier component and a number of sideband pairs that depend upon the modulation index. The carrier and sideband spectra for an fm wave of ±10-kHz deviation and 2000-Hz modulation is shown in fig. 1.

If the magnitude of the fm wave is to be kept constant and at the same time contain a number of changing sideband pairs, it is obvious that the carrier amplitude itself must vary up and down with modulation. In fact, at certain index values the carrier level reduces to zero. The first five such null points are: 2.405, 5.52, 8.654, 11.792 and 14.931.

Spectra distribution for the first two index values are given in fig. 2. A 1000-Hz modulating tone is assumed. You can gain a practical understanding of the technique by considering what happens when a 1000-Hz tone is applied to the input of the fm transmitter and its amplitude is increased gradually from zero to produce higher and higher deviation levels. Obviously there is an increase in deviation and an increase in the modulation index. If this is done gradually and the level of the carrier only is measured, it is found that its magnitude moves up and down. When the amplitude of the 1000-Hz notes rises to a point at which the modulation index is 2.045, the carrier magnitude falls to zero. At this point the deviation would be:

deviation = modulation index x audio frequency

deviation = 2.405 x 1000 - ±2.4 kHz

If you now continue to increase the audio, the carrier once again rises and falls in magnitude. When the index of 5.52 is reached, the carrier again falls to zero. The deviation at this carrier frequency is:

deviation = modulation index x modulating frequency

deviation = 5.52 x 1000 = ±5.52 kHz

As the magnitude of the audio frequency

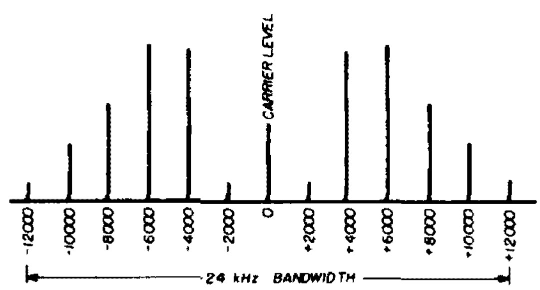


fig. 1. Carrier and sideband distribution for modulation index of 5. Modulating frequency = 2 kHz.

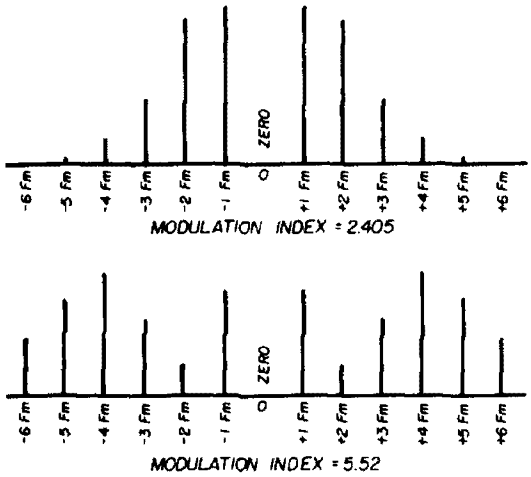


fig. 2. Spectrum distribution for first two carrier-zero positions.

is increased the carrier will pass through a succession of zero positions. For the fifth carrier-zero point (using 1000-Hz modulation), transmitter deviation would be approximately ±15 kHz.

You can use these carrier-zero points to advantage in making carrier-deviation checks. To use them the emission spectra must be displayed on the screen of a spectrum analyzer. Alternately, a frequency meter or receiver could be used if they contain sharply tuned resonant circuits able to delineate the carrier from the first pair of sidebands. The functional block diagrams of fig. **3** show typical test setups. A spectrum analyzer permits the display of carrier and sidebands in accordance with figs. **1** and **2.** Such an oscilloscopic analyzer can be connected directly to the i-f system of an fm receiver. More expensive analyzers that operate in the vhf band can be connected directly to the transmitter output using an appropriate dummy load.

You can make carrier-zero measurements with relatively inexpensive setups. A vhf converter can be used to supply signal to a high-frequency a-m receiver. The S-meter circuit of the receiver must be such that it responds to carrier level only. One of the older a-m receivers with a sharp crystal filter is ideal; I have used the old National NC-109 successfully. Use a high-enough modulating frequency that the first pair of sidebands (1000 Hz and up) does not influence the deflection of the S-meter measuring the carrier amplitude. The S-meter reading should fall off to zero as you tune between the carrier and the first sideband. *Remember that you need a highly selective a-m receiver with an S-meter that responds faithfully to carrier level but not to modulation.* The modern single-sideband receiver cannot be used for this type of measurement because of its S-meter system.

An accurate and highly selective frequency meter can also be used to identify a center-frequency null. Such a frequency meter must have sharply-tuned and high-Q resonant circuits to be able to delineate between carrier and sidebands.

practical audio frequencies

In amateur and commercial fm two-way radio systems, narrow-band (±5 kHz) and wide-band (±15 kHz) deviation are common. Many radio amateurs settle on an in-between value of approximately ±10 kHz so their signals might be demodulated by receivers designed for either narrow-band or wide-band demodulation. Practical tone frequencies of 1000, 1200 and 1800 Hz are attractive for making deviation measurements. An oscillator can be set precisely on frequency by choosing the standard tone transmitted by WWV and a Lissajous pattern display.

The second carrier-zero checkpoint of 5.52 in conjunction with a 1000 Hz note

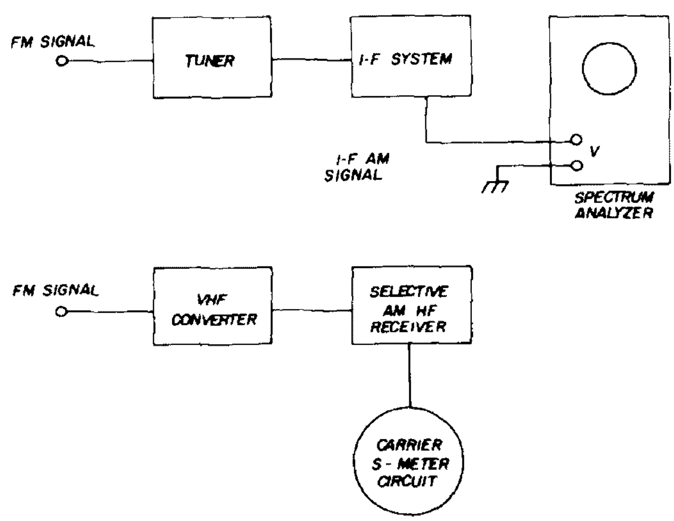


fig. 3. Carrier-zero test setups.

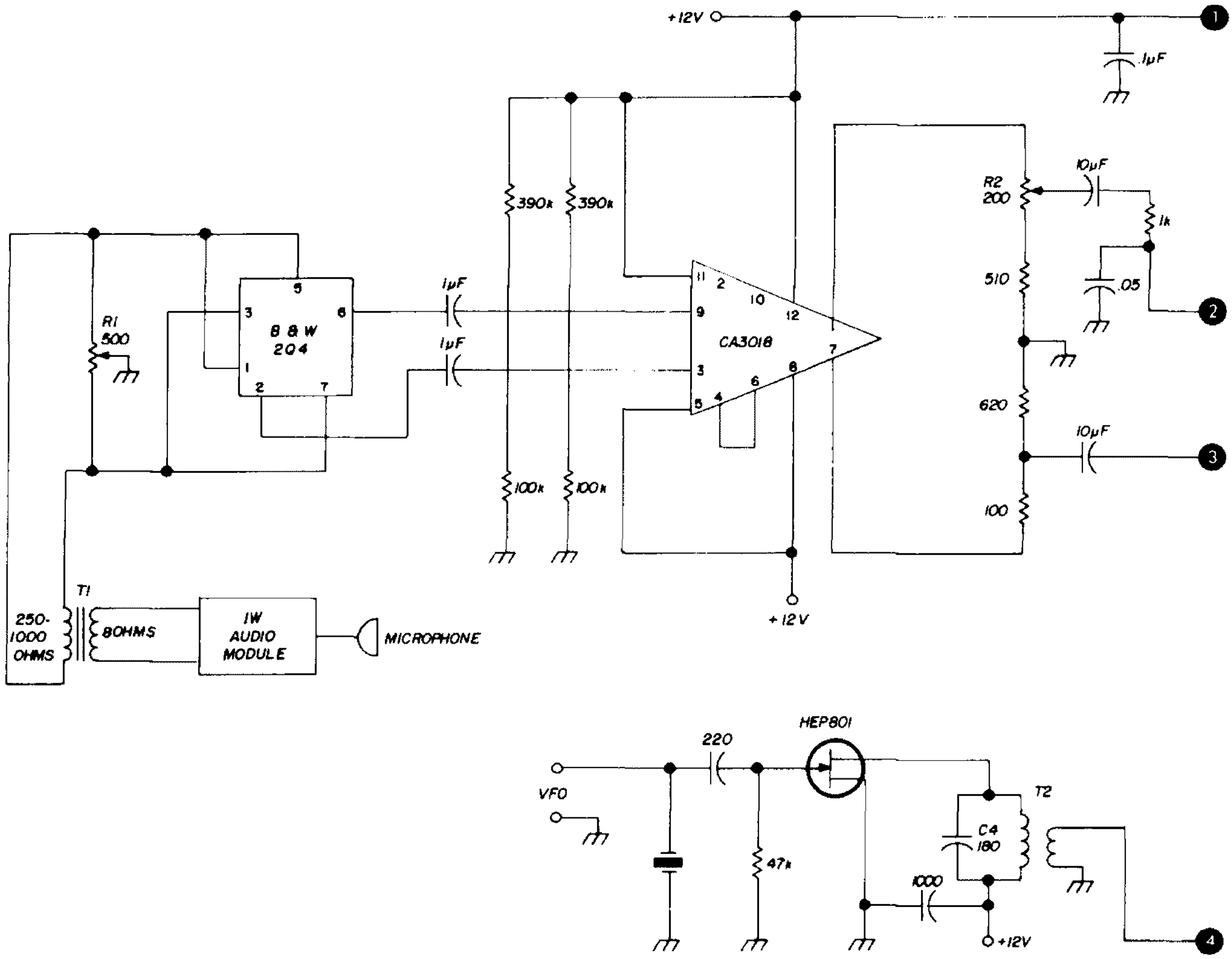


fig. 4. IC phasing-type sideband generator for 160 meters.

can be used to determine the setting of the deviation control needed for ±5.52-kHz deviation. If you now measure the audio-input amplitude to the fm transmitter you will know just how much voice signal is needed for this amount of deviation.

A 1200-Hz tone and the third carrier-zero point is an excellent checkpoint for determining deviation of ±10 kHz.

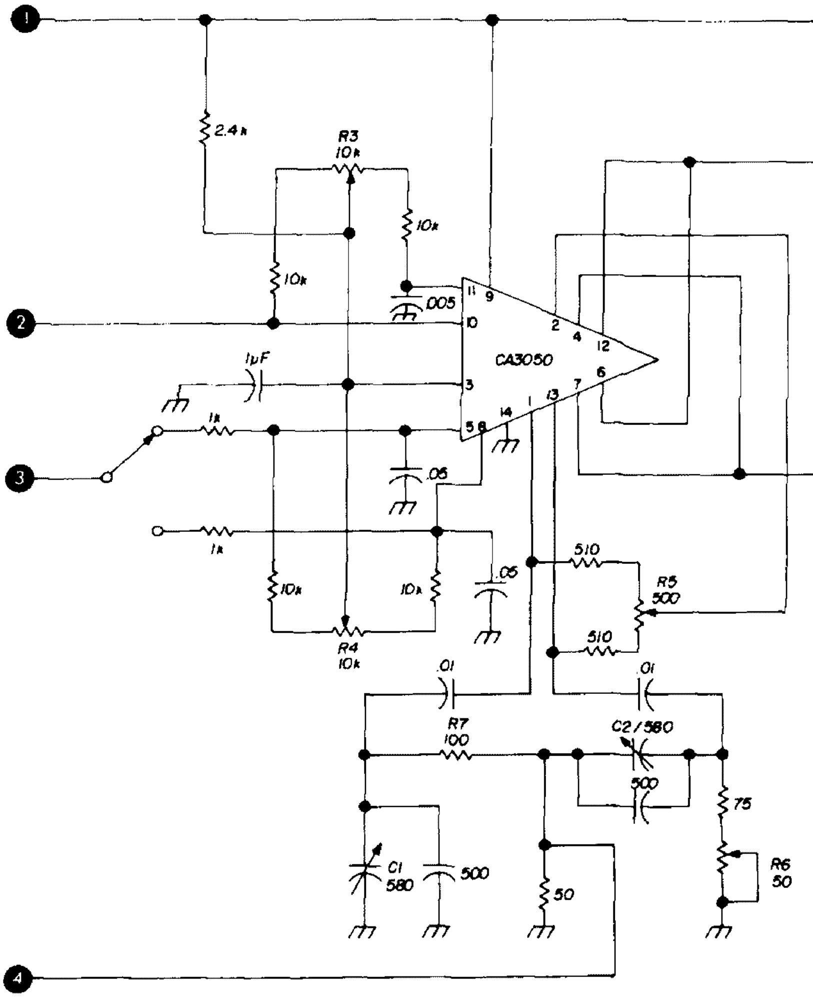
The third carrier-zero point and 1800-Hz tone or the fourth carrier-zero point and 1200-Hz tone can be used as check points for indicating maximum wide-band deviation. Corresponding calculations are:

deviation = 5.52 x 1000 = ± 5.52 kHz
deviation = 8.654 x 1200 = ±10.30 kHz
deviation = 8.654 x 1800 = ±15.58 kHz
deviation = 11.792 x 1200 = ±14.15 kHz

The carrier-zero method of measurement is especially useful in checking out home-built fm transmitters. You can use inexpensive equipment and make use of one of the good old a-m receivers that have been sitting idly around the shack. Most importantly it gives you a better insight into just what takes place in using frequency modulation.

top-band sideband generator

In the August issue[1] I presented a general schematic diagram for a phasing-type single-sideband generator using two RCA integrated circuits. I build a 160-meter version of this generator using the B&W 2Q4 audio phase-shift network shown in fig. 4. The balanced center-tapped primary and low-impedance output winding were wound on a toroid core.



C1,C2	105-580 pF trimmer
C3,C4	10-180 pF trimmer
T1	audio transformer, 250-1000 ohms to 8 ohms.
T2	primary, 80 turns no. 26 enameled, center tapped, on Permacor 57-1541 toroid core. Primary is 15 turns no. 26 over the center-tap area of the primary winding
T3	73.5 µH slug-tuned coil (J. W. Miller 21A825RBI) with 12 turns no. 26 secondary wound below original winding

The generator was chassis-mounted, fig. 5, for ease of adjustment, measurement and self-education. Test instruments were a 1000-Hz tone oscillator, audio voltmeter (fet vom), 160-meter sideband receiver and service-type oscilloscope. The latter is an optional piece of test equipment, but it is very useful and instructive because it can display the various sideband waveforms. I used the 1-watt audio module shown in fig. 4 for voice testing. A transistor output transformer connected in reverse stepped up the low impedance of the audio-module output to the higher input impedance of the phase-shift network.

The rf output transformer T3 was wound on a Permacor 57-1541 toroid core. The primary consists of 80 turns no. 26 enameled wire center-tapped while the secondary is 15 turns of no. 26 overlapping the center-tap area of the primary. The rf oscillator-amplifier transformer T2 is a J. W. Miller (21AB25RBI) coil; the secondary winding consists of 12 turns no. 26 wire on the same coil form below the original winding. The audio transformer T1 is an output transformer with the secondary (8 ohms) connected to the output of the module and the primary (250- to 1000-ohm range) to the audio input of the sideband generator.

A fet oscillator-amplifier stage was also added. It operates as a crystal oscillator, or it will function with the crystal removed as an amplifier and isolation stage for use with an external 160-meter vfo. The carrier signal is applied to the two 45° rf phase-shift networks which supply 90°-related carrier components to terminals 1 and 13 of the CA3050 IC. The combination fixed- and trimmer-capacitor values permit proper 90° relations on the 160-meter band.

tuning

A good first step is to adjust the audio and rf phase-shift networks. The audio-input attenuator R1 is set to midposition. Set potentiometers R3 and R4 to midposition. An oscilloscope or audio voltmeter can be used to observe the 1000-Hz tone at terminals 10, 5 and 8 of the CA3050. Place a 50-ohm carbon resistor across the sideband-generator output. Potentiometer R2 can then be adjusted so the audio level at pin 10 is the same as at pins 5 or 8, depending upon the position of the sideband switch.

The objectives of the rf phase-shift adjustments are to supply equal-level and 90°-related rf components to pins 1 and 13 of CA3050. How do you determine when a 90° phase shift occurs? Forty-five degree networks are involved, and in any series resistor-capacitor combination voltages of equal value appear across the resistor and capacitor when a 45° relationship is established. Your indicating instrument can be an oscilloscope or rf voltmeter that can measure up to 2 MHz.

Initially, the 50-ohm potentiometer R6 is set to midposition. Potentiometer R5 is also set to midposition. Capacitor C1 is adjusted for equal-magnitude levels across R7 and C1. Likewise the trimmer capacitor of the C2 combination is adjusted for equal levels across the C2 and R6 series combination.

Keep the 50-ohm terminating resistor across the output and connect the output of the sideband generator to the antenna input of the sideband receiver. Apply the rf-carrier component only. Turn back the rf gain-control setting or detune the pre-selector so you do not overload the receiver. Adjust the level for an S9 meter reading. Open up capacitor C3 several turns away from its maximum setting. Now adjust the tuning-capacitor C4 for maximum S-meter reading.

Adjust capacitor C3 and potentiometer R8 alternately for minimum S-meter reading. A slight readjustment of potentiometers R5 and R6 may help a bit in further reducing the carrier level. Make these readjustments carefully and be

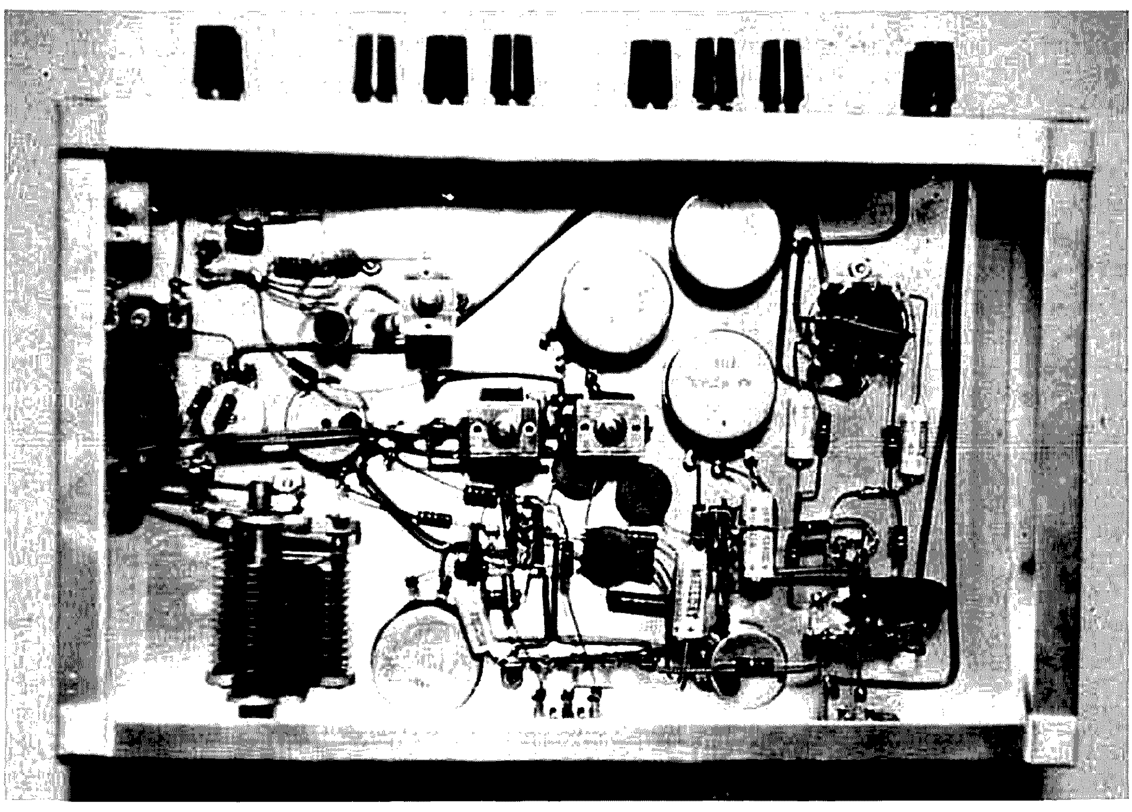


Photo of the ssb generator. The phase-shift network is the socket in the right hand corner.

certain that you can return them to their initial positions if necessary.

You can also make oscilloscopic observations with a high-impedance probe.

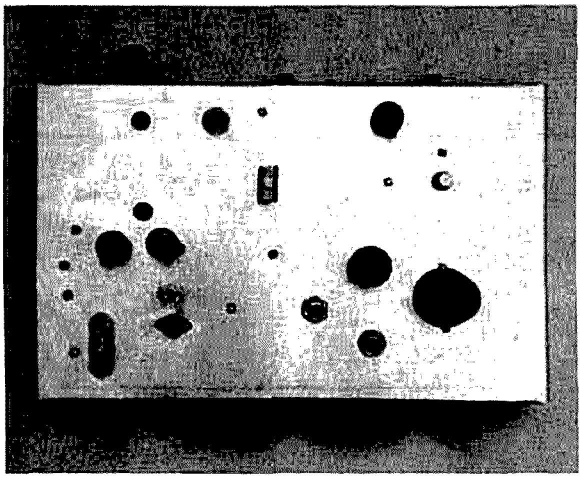


fig. 5. Chassis layout for the 160-meter ssb generator.

Since the probe attenuates the desired signal a problem that often arises is direct carrier pick-up by the scope lead which can result in a false pattern display.

Apply a 1000-Hz audio tone, tune in the modulation on the receiver and check the upper and lower sidebands to determine which position provides the highest ratio between the desired sideband and the undesired sideband. This procedure permits you to determine the position to set your phasing sideband switch.

Throw the phasing sideband switch to the opposite position. Also change your receiver sideband switch to the other position. Your dominant sideband should now have moved from one side of the carrier frequency to the other.

Set up the combination for operation on lower sideband. Now tune to the opposite sideband and adjust potentiometer R1 for minimum undesired sideband. Very slight readjustments of potentiometers R3 and R4 as well as R2 may aid in the reduction of the undesired-sideband level.

reference

1. Ed Noll, W3FQJ, "Circuits and Techniques," *ham radio*, August, 1971, page 50.

ham radio

the ham notebook

color coding parts

Fearing that I would drop a parts bin full of unlabeled surplus diodes and transistors – as Poly-Paks and other bargain assortments often come – I developed a scheme for color coding them. For 15 cents per bottle, I bought a ¼-ounce jar of quick drying enamel for each color of the standard color code. After experimenting with expensive artist's brushes, I found that the lowly round toothpick worked best for marking the components. You can use a different toothpick for each color so you don't have to clean the brush at the end.

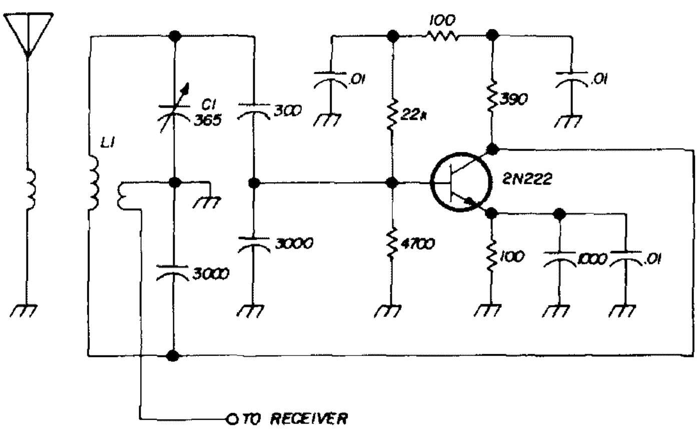


fig. 1. Selective rf amplifier for 80 meters. L1 resonates to 3.5 MHz with 365-pF broadcast variable. Coupling coils (3 turns each) are spaced 1" from L1, not wound around it.

To mark transistors I do not put any mark for the 2N- prefix; I just paint on the last digits. For a 2N706, for example, I put three dots of color on the side of the case – purple, black and blue (706). For an epoxy diode with an 800 volt PIV rating, I paint on a grey, black and brown dot – with the grey dot at the narrow (cathode) end. For power diodes I mark the PIV rating the same way but I put a colored dot on the stud for the current rating – brown for one amp, red for two amps, and so forth.

Paul M. Rich, WA7BPO

selective rf amplifiers

All rf amplifiers and preamplifiers are characterized by broad bandwidth. Because of this they are easily overloaded by static and strong signals several MHz away from the operating frequency. The resultant distortion products from the rf amplifier are fed on to an inherently broadband mixer. The i-f is supposed to clean up the mess, but it can only eliminate those signals that are outside of its passband; the distorted signals within the intermediate frequency are passed on to the detector and audio amplifier.

The answer to this problem is more selectivity in front of the rf stage. This can easily be done by increasing the Q of the input circuit and decreasing the coupling. On 144-MHz this can be accomplished with a high-Q cavity. The cavity is fairly selective, but has some insertion loss, and can be built *after* taking a course

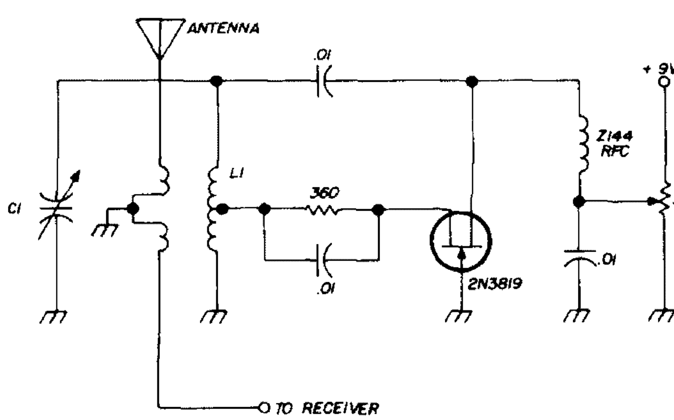


fig. 2. Selective rf amplifier for vhf use. C1 is 35-pF per section split stator variable with insulated shaft (see fig. 3).

in plumbing and silver soldering. The required cavity size for six meters makes it a nice monument to your ingenuity.

High selectivity can also be added to oscillator circuit shown in **fig. 1**. For 144 MHz the fet circuit in **fig. 2** is doing wonders.

Gus Gercke, K6BIJ

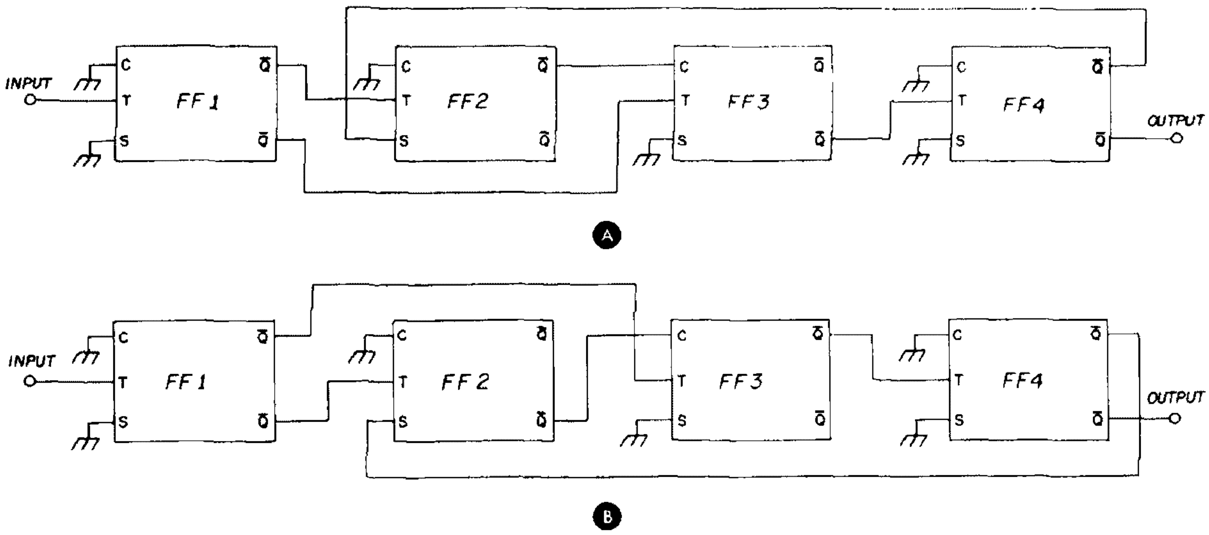


fig. 4. Standard divide-by-10 circuit is shown in (A). In (B) flip-flops 1 and 3 are reversed internally and must be wired as shown to divide by 10.

the rf stage by introducing some regeneration, much in the same fashion as is used in Q-multipliers. This approach will provide insertion gain instead of loss, and you don't have to be a plumber.

An oscillator circuit that is sufficiently stable for use as a vfo can be used as a selective rf amplifier by simply loosely coupling the antenna and receiver input to the oscillator tank coil and adding a potentiometer to drop the supply voltage to the point where oscillation just ceases. Gain and selectivity of this system can be adjusted (one at the expense of the other) by moving the two coupling coils closer or further away from the vfo tank coil. On 80 meters I use the modified Vackar

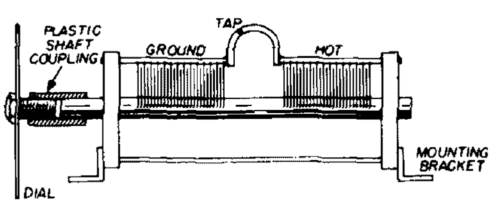


fig. 3. Tank circuit for selective two-meter rf amplifier. Split-stator capacitor is 35 pF per section insulated shaft. For two meters, coil is ½ turn, 1" diameter; six-meter coil is 2 turns, 1" diameter.

using the flop-flip

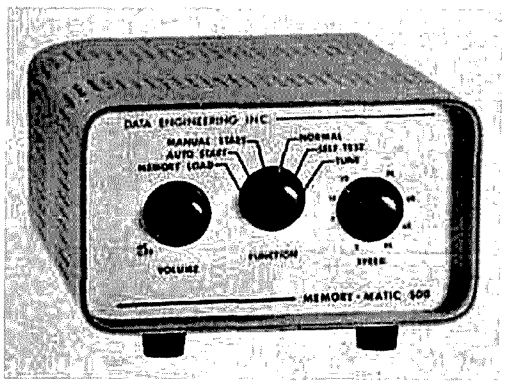
A recent bargain hunt turned up some RTL flip-flops in TO-5 cans; I bought some to make up a logic demonstration board for my sons to work with. Unfortunately, they could not get the modulo-ten minimum hardware circuit to count by ten as it should. After checking the patch connections I put a meter on the flip-flops and found that some of them were actually *flop-flips*.

With the normal RTL flip-flop connected as a binary divider (all inputs but toggle grounded), when you initially apply power Q should be low and $\overline{Q}$ should be high. If some of your bargains are *flop-flips* the reverse is true. I got the modulo-ten circuit to count by ten by treating the normal Q and $\overline{Q}$ outputs of the *flop-flips* as though they were reversed at the pins of the IC. **Fig. 4A** shows the normal circuit if all flip-flops were actually flip-flops. **Fig. 4B** shows a circuit where flip-flops 1 and 3 are *flop-flips*. These inverted ICs are usable in this fashion once you understand their peculiarities.

Allan Joffe, W3KBM

new products

ic keyers



The Memory-Matic 500 and the Space-Matic 21 are two new electronic keyers from Data Engineering. They were developed especially to meet the demands of professional and amateur code operators. These keyers were designed to send *all* code elements: dot, dash, dot space, dash space, character space and word space in an instant-start, self-completing mode with a guarantee of no missed or extra dots or dashes and provides jam-proof dot, dash, character and word spacing. Each code element is automatically generated with little or no effort on the part of the operator.

The memory system included with each Memory-Matic 500 provides the serious DX, contest or traffic operator with provision for instant storage of code characters or messages, for an immediate reply to on-the-air contacts.

Messages of approximately 40 code characters are easily keyed into the memory for calling stations, giving a contest exchange, calling CQ or testing. Loading code characters into *memory* or transmitting messages from memory is accomplished at the same speed and weighting ratio in use by the operator. The memory allows continuous transmission on a repetitious basis of any message in storage.

Iambic squeeze keying is provided through the use of twin paddles. Iambic operation provides alternate dots and dashes when both paddles are squeezed. Advanced dot-dash memories automatically insert a dot in a series of dashes or a dash in a series of dots and insures against missed or extra dots or dashes.

Jam-proof spacing is provided for dots, dashes, characters and words. Dot and dash spacing is automatically added following each dot or dash. Character spacing is automatically added following the last dot/dash element which make up each character. Word spacing is automatically added following the last character which makes up each word.

A 500-bit instant-load memory is provided in the Memory-Matic 500 which can be programmed or updated by moving the function switch to *memory-load* and keying any message into memory. When 50 storage positions remain in memory the monitors's sidetone pitch automatically increases, indicating to the operator that the memory is *near-full.* The monitor will emit a steady sidetone if the memory overflows; this condition necessitates reprogramming of the memory-load sequence. Once a message is entered into memory it can be transmitted manually or automatically at intervals ranging from a fraction of a second to several minutes. Adjustment of the automatic transmission interval is accomplished by a *rear panel control.*

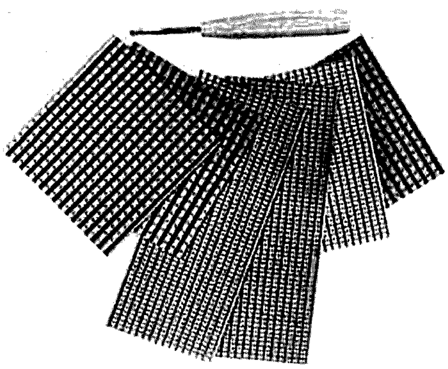
Both the Memory-Matic 500 with the 500-bit memory and the Space-Matic 21 without the memory feature have a speed range of 5 to 85 words per minute with adjustable weighting, independent of speed. They have provision for keying

More Details? CHECK–OFF Page 94

with a regular straight key and keyer activation by either a single or dual paddle key. Tune switch, variable pitch sidetone oscillator, 117-Vac power supply, and off-the-air self-testing circuitry are built into both keyers.

The units come with instructions, all cables and connectors and are guaranteed for a year. The Memory-Matic 500 costs $198.50 and the Space-Matic 21 is $89.50. Use *check-off* on page 94 for more information or write directly to Data Engineering, Inc., Box 1245, Springfield, Virginia 22151.

veroboards



The Vero method of construction does away with the etching, drilling, and tools usually associated with breadboarding. A perforated board has a series of parallel conductor paths connecting all the holes in one column. After inserting component leads in appropriate holes, a scriber is used to break the conductor paths where they are not needed. Where paths are needed but are not provided by the foil, the component leads may be used, jumpers inserted or special pins inserted in the holes for tie points.

Vero offers an introductory BK-6 Vero-board kit designed for experimenting with discreet components.

The BK-6 Kit consists of six Veroboards, two with a 0.2 x 0.2-inch matrix and four with a 0.156 x 0.1-inch matrix, both having a 0.052 diameter hole. The Vero BK-6 Kit sells for $5.95.

Vero also offers the new MC-10 kit for dual in-line integrated circuits. ICs can be plugged in anywhere on the 0.1 x 0.1-in. drilled matrix.

The kit contains a single sided 18-inch long Veroboard with 34 copper strips on 0.1 centers, a plug-in single-sided board with 32 plated contacts, a double-sided plug-in board with strips at right angles and an epoxy glass plain board pierced at 0.1 centers with 0.052 holes. There are also 500 terminal pins to fit the 0.052 diameter holes in a 0.1 plain Veroboard, a pin insertion tool, a spot face cutter to break the copper strip where required, a 0.1 reproducible design sheet and a 32 contact 0.1 edge connector. The Vero MC-10 Kit sells for $29.95.

For more information write to Vero Electronics, Inc., 171 Bridge Road, Hauppague, New York 11787 or use *check-off* on page 94.

new callbook

Marking it's fiftieth anniversary, the Radio Amateur Callbook Magazine is changing to a new format of one major edition per year. This first of the new Callbooks will be the Winter 1971-72 issue, published on December first.

A "new information service," in Callbook form, consisting of new licenses, silent keys, call letter and address changes for the preceeding three months, will be initiated on a quarterly basis — every March 1, June 1 and September 1 — to those who have purchased the previous December issue. This information service will be available by subscription only, through the order form printed in the December edition. The price for this service will be $6.00 per year for the United States series, and $4.50 per year for the foreign series. The subscription is on a consecutive, annual basis only.

The Winter U. S. Callbook and three supplements will cost $15.20 per year, and the Foreign Callbook and supplements will cost $11.70. The U. S. Callbook alone will remain at $8.95; the Foreign Callbook is $6.95.

For more information use *check-off* on page 94 or write to the Radio Amateur Callbook Magazine, 925 Sherwood Drive, Lake Bluff, Illinois 60044.

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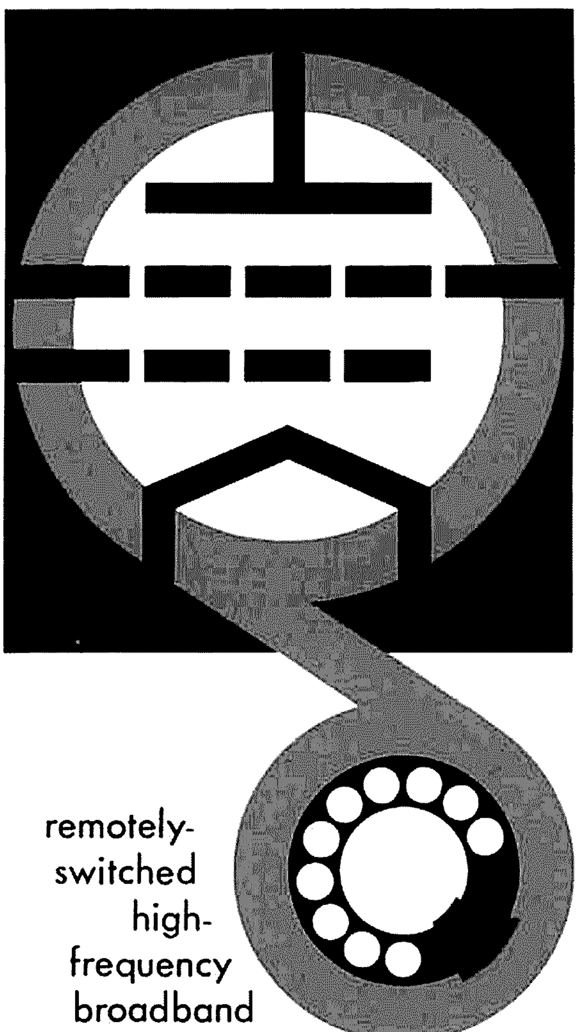
MARCH 1972



remotely-switched high-frequency broadband

linear amplifier

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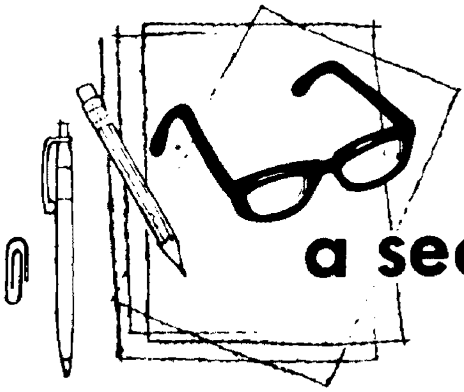


a second look

by jim fisk

During the **5th annual SAROC** convention in Las Vegas in January, A. Prose Walker, W4BW, new Chief of the FCC Amateur Division, held forth for more than an hour, offering many of his personal views on important aspects of amateur radio.

On the current repeater proposal before the FCC, he favors minimal legislation. He feels that repeaters should only be linked together when it is justified in view of better emergency or public safety performance; linking repeaters solely for better range is absurd. Rather than tying up a number of repeaters and frequencies, it is much better to use a lower frequency. In his view, when repeater linking *is* allowed, it will be limited to two or three stations, and then only for a good reason.

As far as the new expanded phone bands are concerned, Mr. Walker said he feels there will be a token expansion of the 75-meter phone band. The 40-meter phone band is likely to be expanded too, but not enough to cover the inter-zone DX window below 7.1 MHz. In his opinion, expansion of the 20-meter phone band "will cause too much international ill-will and at this time we need all the friends we can get." He spoke of this in the context of what is in the best interest of the United States, not just amateur radio. He was very strong on this point, much to the chagrin of some of the big DXers in the audience.

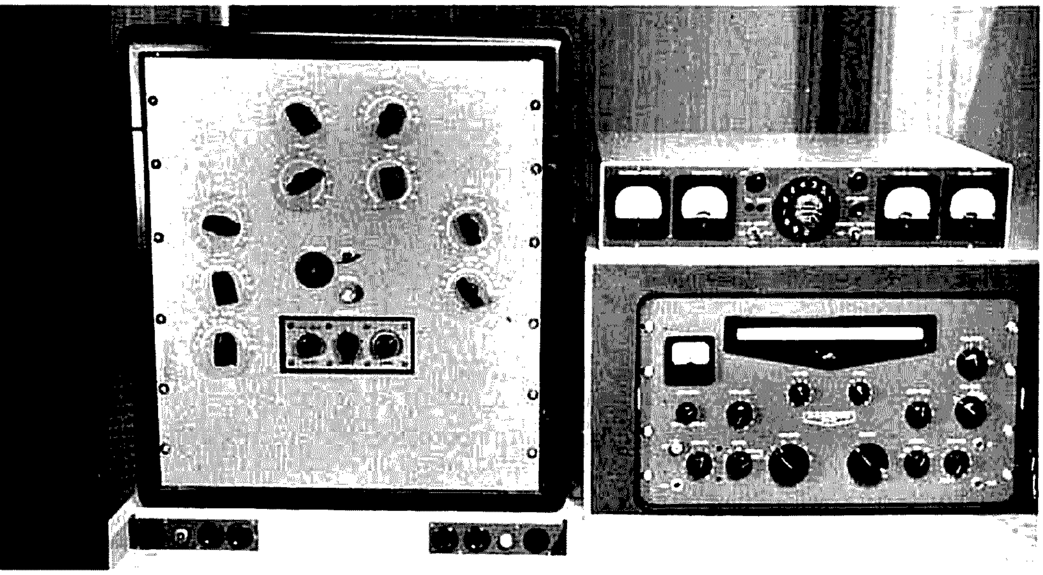
Walker pointed out that running more than a kilowatt is illegal, unsportsmanlike and totally out of keeping with amateur tradition and the purpose of the amateur service. You know who is doing it, and the FCC knows. He advised amateurs to clean up their own house – or the FCC will do it well and do it painfully. His personal opinion is to prosecute to the fullest extent of the law.

Looking to the future, Mr. Walker suggests that we may not have to identify; all licensed stations will have *built-in* electronic identifiers which will do it automatically, at very high speed. This will automatically, and positively, identify jammers and bootleggers.

Far in the future he feels our whole callsign structure may be changed with distinctive calls for each class of license. There are enough letter combinations for all generals to have a one-by-three call (such as W1AAA). Higher classes of licenses may use calls like W1A, WA1A, W1AA, etc. There is also the possibility that the ITU-allocated block of U.S. calls from AAA to ALZ might be used for the amateur service, if and when Tibet, Sikkim and Bhutan amateurs stop using their AC1-ACØ callsigns. This is far in the future, though.

There's also a long-range possibility for more bands as more and more commercial and military traffic switches to land lines and vhf. Then the real frequency pressure will come on vhf and uhf. And, in light of the EIA proposal for 220 MHz, W4BW encouraged hams to, "Use it, or lose it!" That goes for all of our bands.

Jim Fisk, W1DTY
editor



ecology linear

This high-power linear amplifier eliminates air-wave pollution of on-the-air tuneup with broadband circuits and nearly instantaneous band changing

Gene Hubbell, W7DI, 6633 East Palo Verde Lane, Scottsdale, Arizona

Waste, pollution, ecology – words much in evidence today, and meaningful in amateur radio, as well as in our living environment. For years I have, occasionally, blown my top over unnecessary signals that were "dirtying up" the bands. There seemed to be a quota of fellow hams who needed to test or tune up regularly on "my" frequency. Of course, many of them felt the same way about me.

Some time ago I adopted a tuning procedure that cut my own on-the-air tuning to a minimum. Even though I still make a final check with the antenna connected, I minimize this by setting all dials to recorded values, using a dummy antenna up to the last moment, and picking a test frequency carefully to avoid interfering with communications.

If all controls could be set for each band, and a rapid change made between preset circuits, I would not only avoid putting out unnecessary signals, I'd have an ideal transmitter for contest work. Of course, separate finals would do this, but that solution seemed a lazy man's way out. Besides, the space they required would crowd my shack. Remote control

would save space in the shack if space were available elsewhere; it would be worth remembering.

After reading through many magazine articles on high-power linears, considerable doodling on paper and shuffling of parts on the workbench, I started to build. Some 800 hours of work later I had the linear amplifier shown in **fig. 1**. Input power runs from zero to well over 2 kW, complete band changing takes seven seconds or less. It can be switched from band to band by a local control on the front panel, or by a telephone dial on the remote-control unit. The plate supply is continuously variable from zero to 5000 volts, can be left on continuously or automatically keyed on and off by an auxiliary set of contacts on my antenna relay. Or I can control it with a vox relay. Time-delay circuits provide for gradual application of primary power to avoid surges, and also to hold the plate supply on during pauses in talking or keying.

the circuit

When planning the plate tank circuits I decided that only vacuum capacitors would fit into the desired space, so four 10- to 300- or 400-pF variables were acquired with a fixed unit of 12 pF planned for the 28-MHz tank. The inductances were wound of 3/16-inch copper tubing, except for 28 MHz, where small copper strap was used. The output coupling air variables were 30- to 500-pF units found in military surplus. The 14-MHz variable was shortened slightly to avoid touching the big tube.

On the three lower bands fixed mica capacitors shunt the air variables to provide a 50-ohm output impedance. All plate circuits were designed for a Q between 10 and 15, depending on the band and the plate voltage I planned to use.

The large ceramic band-switch decks were assembled with mycalex and bakelite spacers made up on a lathe, so there were no closed loops in switch construction. The turning mechanism for the plate tank band-switch was originally a Collins auto-tune. This motor-driven gear and clutch assembly ran from 117 volts ac and would turn a shaft to ten positions. A small control panel attached to the auto-tune had a 10-position switch to select these positions, each of which was independently adjustable over 360 degrees. A "local-remote" switch made these positions available through screw terminals which were connected to a stepping switch.

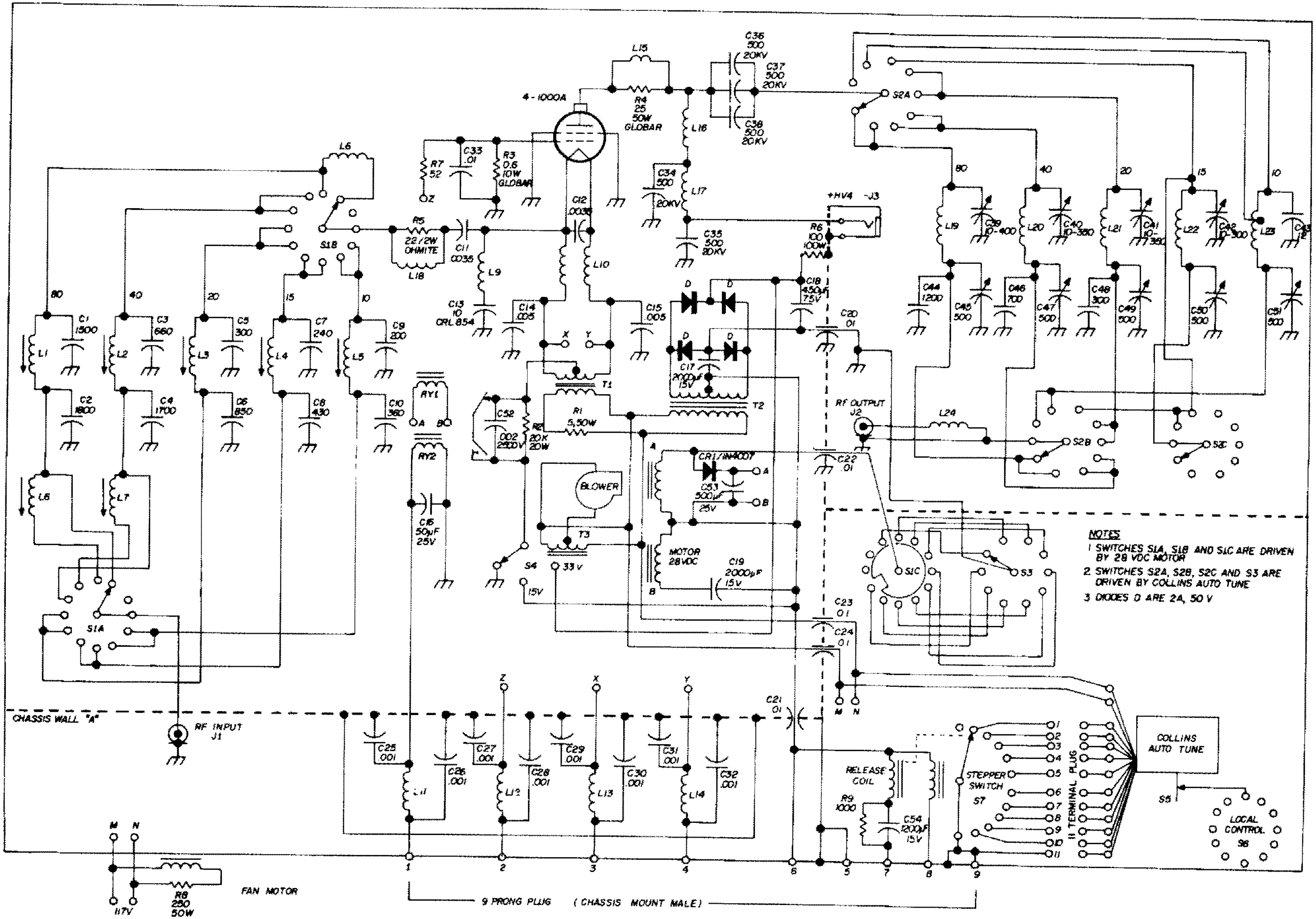
Another set of switch contacts made it possible to switch another circuit to 10 different connections and was used to control another motor-driven coil switch. This item was obtained from an old military receiver.

The reason for the separate band-switching motor drives lay in the fact that the plate tank switch had 10 positions (9 active and 1 rotor) while the cathode tank switch had 12 positions. A suitable 5-to-6 gearing ratio was not readily available.

The cathode drive circuits were designed for low Q with slug-tuned coils wound of number-18 wire on ¾-inch diameter ceramic forms. Input and output capacitors are small silver micas, paralleled for needed capacity and also to carry higher current. Series inductances on other switch connections broadband the coverage on 3.5 and 7 MHz.

All tank circuits were initially adjusted by using a small 10-watt vfo-controlled transmitter with a Micro-Match swr bridge in the 50-ohm line to the circuit under test (input line to the cathode and the output line from the plate circuit). The cathode-to-ground impedance was simulated by a 100-ohm non-inductive resistor; the plate impedance was simulated by a group of resistors totalling 5000 ohms. L and C values were changed to obtain as low swr as possible. Other changes had to be made later when power was applied, but this initial procedure was very helpful.

The linear chassis included, in addition to the 4-1000A socket, a blower, filament transformer, input and output tanks and small dual-voltage dc power supply with


4-1000A
RF INPUT J1
RF OUTPUT J2
+HV4 -J3
CHASSIS WALL "A"
BLOWER
MOTOR 28VDC
COLLINS AUTO TUNE
11 TERMINAL PLUG
STEPPER SWITCH S7
RELEASE COIL
LOCAL CONTROL S6
FAN MOTOR
117V
R8 250 50W
9 PRONG PLUG (CHASSIS MOUNT MALE)
NOTES
1 SWITCHES S1A, S1B AND S1C ARE DRIVEN BY 28 VDC MOTOR
2 SWITCHES S2A, S2B, S2C AND S3 ARE DRIVEN BY COLLINS AUTO TUNE
3 DIODES D ARE 2A, 50 V

nominal 15- and 30-volts output. The 15-volt supply is used for the cathode tank switch drive motor, the stepping switch, control relays for the plate supply, pilot lights on the control box, relay power for the cathode relays K1 and K2 and bias for the 4-1000A. The 30-volt output is also available for bias, and three values, zero, 15 or 30 volts, can be switched in by S4.

fig. 1. Broadband linear amplifier provides rapid bandswitching and variable output from zero to 2 kilowatts.

C39	10-400 pF, 10 kV vacuum variable capacitor
C40,C41	10-350 pF, 10 kV vacuum variable capacitor
C42	10-300 pF, 19 kV vacuum variable capacitor
C43	12 pF, 20 kV fixed vacuum capacitor
L1	12¼ turns no. 18, 7/16" long, wound on ¾" slug-tuned form (National XR72)
L2	6¼ turns no. 18, ¼" long, wound on ¾" slug-tuned form (National XR72)
L3	4¼ turns no. 18, 5/16" lnog, wound on ¾" slug-tuned form (National XR72)
L4	3¼ turns no. 18, ¼" long, wound on ¾" slug-tuned form (National XR72)
L5	3¼ turns no. 18, ½" long, wound on ¾" slug-tuned form (National XR72)
L6	22 turns no. 24, ¾" long, wound on ¼" slug-tuned form
L7	9 turns no. 18, 5/16" diameter, ¾" long
L8	12 turns no. 18, ¼" diameter, 5/8" long
L9	5¼ turns no. 14, 3/8" diameter, ½" long
L10	Bifilar filament choke, 30A (William Deane, 8831 Sovereign Road, San Diego, California 92123)
L11,L12	20 μH vhf choke
L13,L14	
L15	1 turn no. 12, ¾" diameter 3" long
L16	90 μH, 2 amperes (William Deane)
L17	rf choke (Ohmite Z-50)
L18	3 turns no. 18, 1" long, wound on Ohmite 2-watt resistor
L19	17 turns 3/16" copper tubing, 4-3/8" diameter, 5¾" long
L20	14 turns 3/16" copper tubing, 3¼" diameter, 4¾" long
L21	13 turns 3/16" copper tubing, 2-3/8" diameter, 4½" long
L22	9 turns 3/16" copper tubing, 2¼" diameter, 3½" long
L23	5 turns copper strap, 3/8" wide, 0.10" thick, 2¼" diameter, 2¾"long
L24	2 mH rf choke

With zero bias the 4-1000A drew plate current whenever plate voltage was applied. This gave rise to thermal emission noise which fed back to the receiver through the break-in system. Bill Orr advised that a small amount of fixed bias would not greatly affect the linearity, so the bias switching was added. With added bias drive requirements increased slightly.

Operation of the amplifier for initial tune-up is quite simple. With plate voltage off, the local control switch on the front panel, or the telephone dial on the control box, sets the auto-tune in motion to select the proper plate-tank circuit. This also activates the cathode-tank selection switch which locates the correct tank circuit. While this is going on, a maximum of seven seconds, the exciter is tuned up on the proper band segment, and as soon as cathode relay K2 closes, excitation is applied.

A dummy load is switched to the output of the amplifier and with excitation removed, the plate voltage is switched on and adjusted to a medium value of 2000 to 2500 volts. Next, with excitation applied, the plate tank input and output capacitors are adjusted for maximum output to the dummy load. If the plate current appears to be normal (250 to 300 mA) and the output is in the 200- to 400-watt range, the plate voltage is raised to maximum and the capacitors adjusted for maximum output.

With excitation and plate voltages turned off, and the receiver in operation, the antenna is switched in place of the dummy load and a tune-up frequency is chosen. Then excitation and plate power are applied and final tune-up made to the antenna. This usually requires about two or three seconds.

Once these adjustments are completed it is rarely necessary to re-tune unless an antenna change makes a considerable difference in loading. During the 1971 ARRL DX Contest, CW division, I changed bands at least 87 times according to my log, and not once did I have to

tune up "on-the-air." Many times I wished that other stations could say the same.

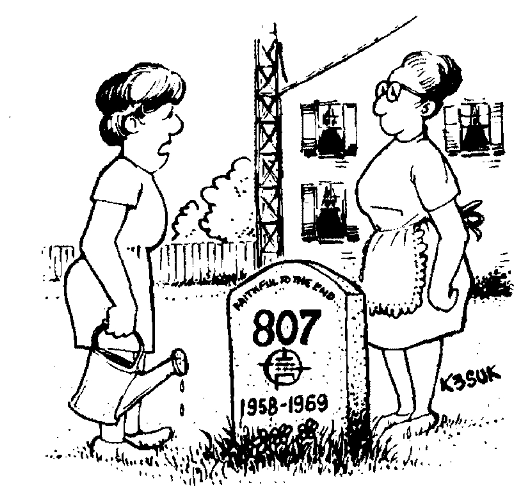
construction

A number of problems turned up during construction. While some required considerable effort to solve, others were handled with ease. One problem that showed up early was the "case of the too-large blower motor."

I had two blowers on hand; both were large affairs with four impellers each, made to cool a large cabinet. I cut one down to a single impeller but it still required 2.5 amperes at 117 volts ac. It blew a gale but was also very noisy; a large part of the noise was from ball bearings.

The second blower with sleeve was much quieter but still drew too much power from the line. Reducing the line voltage gave good air volume and much less power consumption, so I made an auto-transformer from an old filament transformer, teasing up the end turns on the primary until I found the right layer to tap. At 70 volts the motor only drew a little over 1 ampere.

The parasitic suppressor consists of a 25-ohm, 50-watt Globar resistor; copper end caps were forced on and a 3-turn coil of number-12 wire was wound from end to end. This did not completely suppress a parasitic around 130 MHz so a small series LC trap was installed from



"I sometimes think Henry takes his hobby too seriously."

cathode to ground. This is L9C13 on the circuit diagram. Later R5L18 was added in the cathode drive lead. This enabled a reduction in the size of L15 from 3 turns to 1 turn and decreased the power dissipated in R4.

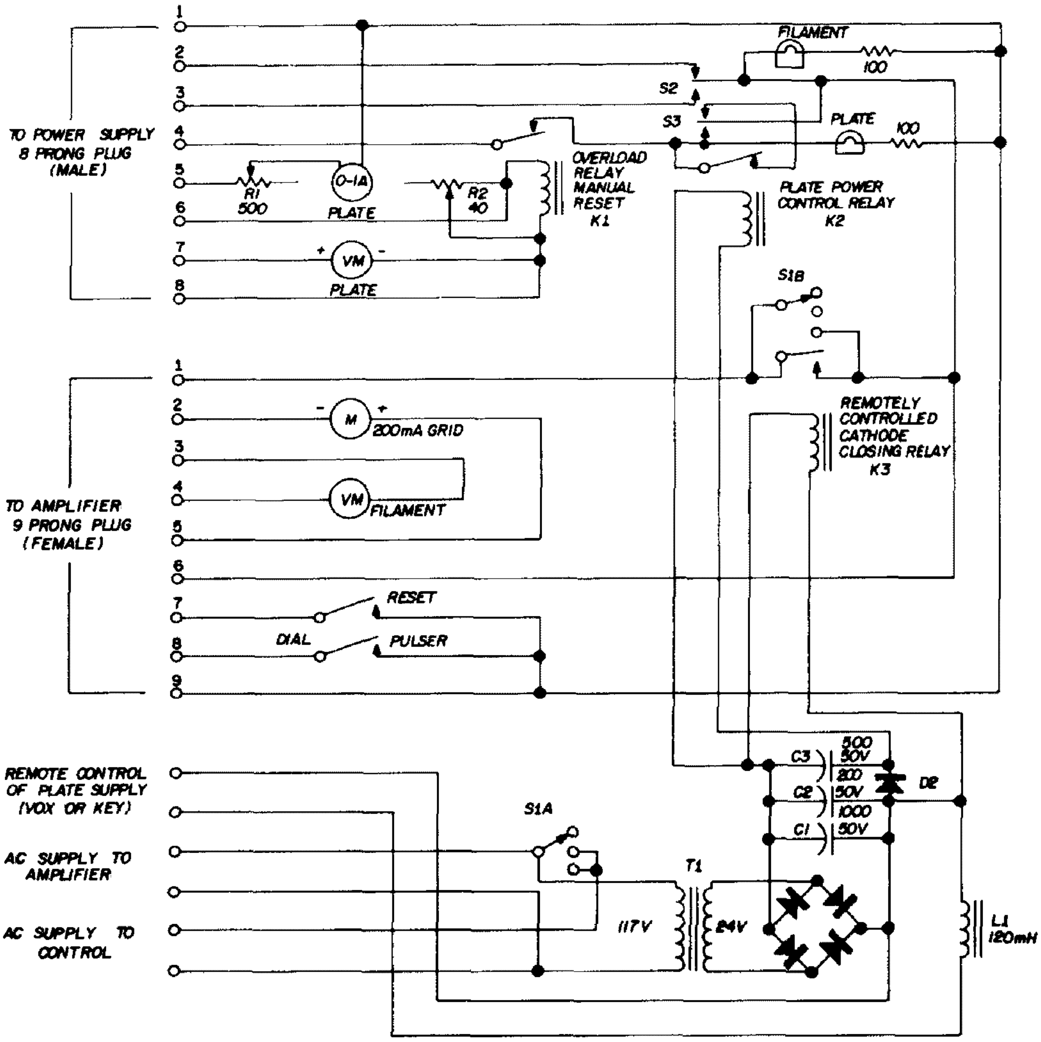
The 21-MHz tank circuit resonated fine on that band with the tube output capacitance, but resonance increased to 28 MHz when the tube was switched to the normal 28-MHz tank. No amount of tinkering with LC ratios, coil sizes or orientation seemed to keep these tanks from coupling with one another, and there was not enough room for shielding, so I resorted to a brute force solution; I added another switch deck to short out part of the 21-MHz coil when operating on 28 MHz; this effectively detuned it.

power supply

The power supply was designed for continuous duty without worry. The chassis was made from ¼-inch thick aluminum plate, 33½ by 14 inches, with vertical 2½ by 1½ sides of 1/8-inch aluminum channel. Thirty-two casters of the type used under refrigerators gave this chassis some mobility.

The plate transformer is rated at 4400 volts at 1 ampere with 230 volts input, and is fed from a 2.7 kVA variable transformer with an output from zero to 260 volts. A pair of 4- to 16-henry, 500-mA swinging chokes are used ahead of a 25-μF, 6250-volt oil-filled capacitor. The chokes were tried in parallel, but the series connection gave far better voltage regulation, and total plate and bleeder current runs only a little over the 500-mA rating.

The bleeder is made up of four 240-watt, 10,000-ohm resistors in series. A time-delay circuit powers a relay to short out a 12-ohm resistor in series with the line voltage to the plate transformer. A small fraction of a second delay allows the input power to be applied gradually, eliminating surges that might trip the circuit breakers or blow fuses or rectifiers. The bridge rectifier band consists of forty 2 amp, 1000-PIV



K1	spst, normally closed, plug-in relay, 8000-ohm coil	T2	filament transformer, 117 Vac primary, 25.2 Vct, 1 A secondary
K2	spst, normally open relay, 12-Vdc coil	T3	117-70 Vac autotransformer
T1	filament transformer, 117 Vac primary, 8 V, 21 A secondary		

fig. 2. Power supply for the high-power broadband linear amplifier. Each leg of the bridge consists of 10 diodes (2 A, 1000 PIV), 10 capacitors (0.0047 μF) and 10 resistors (470k, ½ watt). Plate and filament lamps are 6-volt, type 47.

silicon diodes, 470k, ½-watt resistors and 0.0047-μF capacitors.

The variable transformer in the plate supply primary is driven by a small Bodine geared-head motor with a 30-pm output shaft. Lamp L1 is included to provide motor protection when stalled at the end of rotation and also reduces the speed somewhat. There is some advantage in fast rotation, since it only takes a couple of seconds to shift from maximum to minimum plate voltage. Just a light tap on S2 on the control panel changes the plate voltage by 500 volts or so.

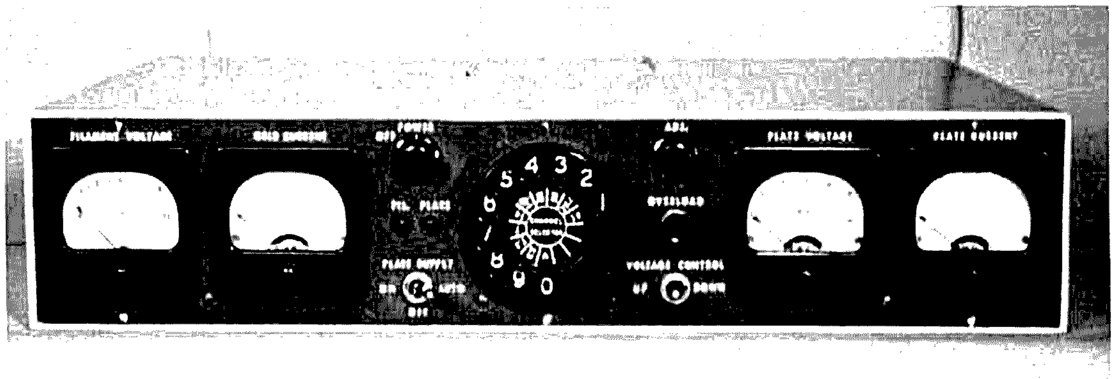
Metering of four items of current and voltage, remotely and safely, poses something of a problem. The ac filament measurement is simple, and luckily, the voltmeter error just about compensates for the voltage drop across the bifilar cathode choke. The plate voltage, plate current and grid current are all measured by setting up a small voltage drop in a part of the circuit to be checked, not much above ground, and calibrating a voltmeter to read in terms of the required units.

The plate voltage is read across a

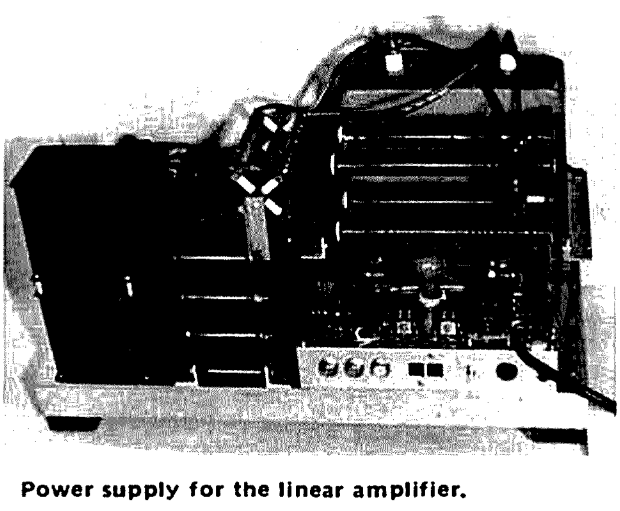
portion of a 50-watt, 300-ohm potentiometer in series with and at the ground end of the 40,000-ohm bleeder string. The voltmeter actually reads about fifteen volts full scale, but the meter dial is calibrated to read to 6000 volts. By varying the potentiometer tap, the readings correspond accurately with a portable voltmeter temporarily connected across the power-supply output.

The plate-current meter is calibrated to read 1000 mA dc, but actually measures the drop across a 1-ohm, 25-watt resistor in series with the negative high voltage to ground. The grid current is measured as a voltage drop across a 0.6-ohm, 10-watt Globar resistor from grid to ground. A 12-ohm, 20-watt resistor in series with the high voltage to ground and the 1-ohm resistor mentioned above develop enough voltage to trip the overload relay on the control panel if the current rises much over 500 mA, so an adjustable shunt was provided for higher current, if desired.

The apparent extra leads in the metering circuits were found to be necessary. A common lead between two metering circuits caused a lot of trouble. High voltage was carried by a length of RG-11/U coax cable with modified connectors. The bakelite inserts on the PL-259 and SO-239 connectors were removed and the center conductor of the coax was allowed to protrude about three inches through the joined connectors, along with its polyethylene jacket and a piece of vinyl tubing to add stiffness. A banana plug was soldered at the end of the coax center conductor. A heavy bakelite tube was added to the inner, unthreaded, side of the SO-239 and a banana jack was installed at its far end. The high-voltage connection was made here; the shielded coax jacket served as ground and the high-voltage negative lead.



Amplifier remote control and metering unit.



Power supply for the linear amplifier.

control

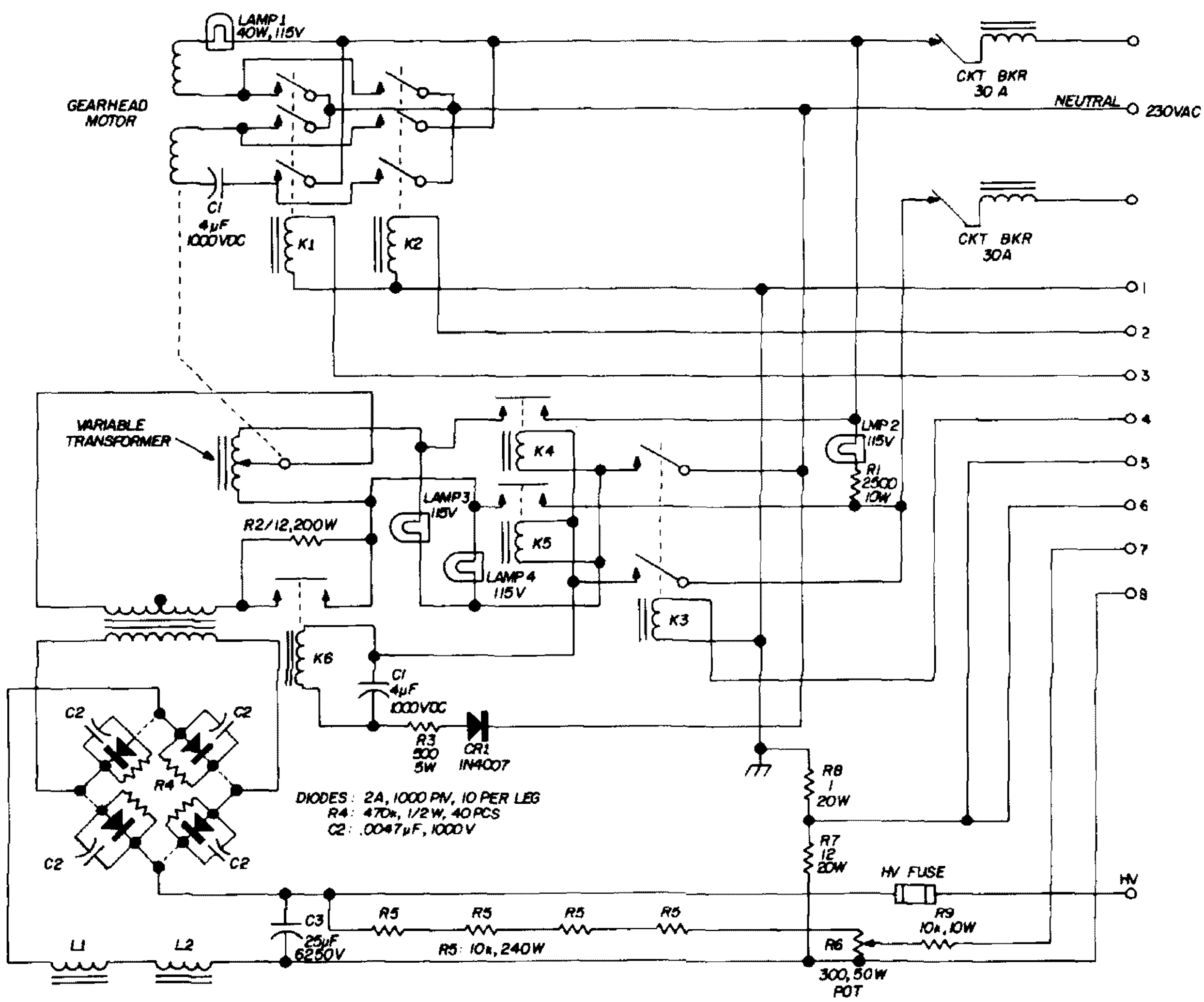
The control circuitry for energizing the high voltage supply was originally designed to operate from the same 15-volt supply which operated most of the relays in the system, closing K2 in the control box. It was desired that K2 and K3 close very quickly and open slowly, at different time delays, and eventually a separate power supply was added. This supply, from T1 in the control box, provided a no-load voltage of around 40

volts with a 1000-μF capacitor for energy storage.

When an external circuit was closed, this voltage was applied to K3 with a 200-μF capacitor across its coil, and to K2 with a 500-μF capacitor across its coil, but with a series diode *preventing* discharged more slowly, thus keeping the plate supply energized for a somewhat longer period of time. The delays are about 1½ and 3½ seconds, respectively.

construction

The linear cabinet was picked up at a

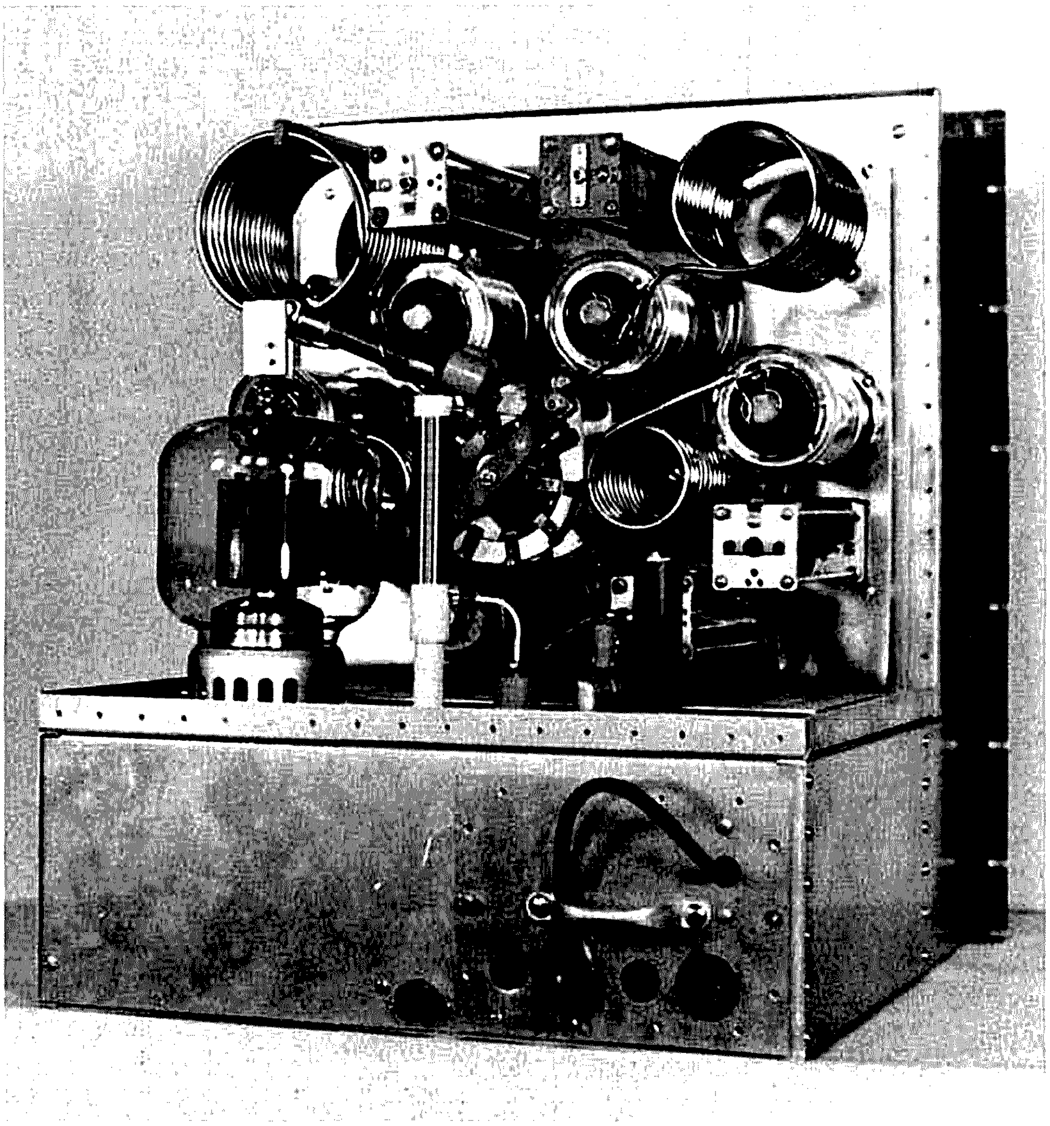


K1,K2, K3	triple pole, single throw relay, 24 or 12 Vdc coil	K4,K5, K6	dpst relay, 30 A contacts, 117-Vac coil
		L1,L2	4-16 henry swinging choke, 500 mA

fig. 3. Schematic diagram for the control unit. Diodes in the bridge circuit are rated at 2 amps.

discharge back through the coil of K3. A small inductance cut arcing at the closing contacts at the remote control point. Discharge of the 200-μF capacitor was relatively fast, allowing K3 to open soon and in turn release the cathode grounding relay, K2, on the main chassis. The 500-μF capacitor across K2 (control unit) surplus store and has outside dimensions of 21½ x 24 x 18½ inches. The inner cabinet and chassis were fabricated of sheet aluminum from a local junkyard, cut and bent on machines in a nearby tin shop. Not counting the screws holding internal components, there are at least 350 machine screws holding the shielding

together. The control-unit cabinet was built of wood, painted to match my console. The labeled front panel of the linear was put on top of the actual support panel, which held many support and mounting screw heads, all recessed. The cabinet and front panel were painted with a two-tone metallic blue-gray at a local auto body shop.



Rear view of the ecology linear shows separate tank circuit for each band and battleship construction throughout.

Other little items that come to mind, which may be of interest, include the over-travel on the cathode tank switching motor. Even though the 28-volt motor was operating on about half voltage and the drive was through a Geneva gear, the drive sometimes ran past proper contact on the switch decks. This was cured by using a dynamic brake circuit with capacitor C19 in series with winding B of the motor. Now the drive system stops – but quick.

To obtain added frequency coverage on 28 MHz with the fixed capacitor C43, I used a tap just off the end of the coil, hooked to another plate-tank switch tap. This raised the resonant frequency nicely, from about 28.1 up to 28.6 MHz.

Relay K1 on the linear chassis was added to open the cathode circuit to ground (except for R2) when the cathode-switching motor started to operate and to close the circuit after the motor stopped, with time-delay capacitor C53.

All plate-tank circuit components were mounted on an aluminum sheet bent at right angles, parallel to the front panel and bottom chassis. This support is entirely insulated from the rest of the assembly except for one ground connection at the tube socket and the braid on the output coaxial cable feed line.

Connections were provided for high voltage, rf input and output, 117 volts ac and a 9-wire control cable socket, all in a 5 x 7 x 2-inch chassis, inside the rear surface of the main chassis. A cover over this opening was drilled for the coaxial cable connectors, and has male plugs mounted on it for the ac line and 9-wire cable connections.

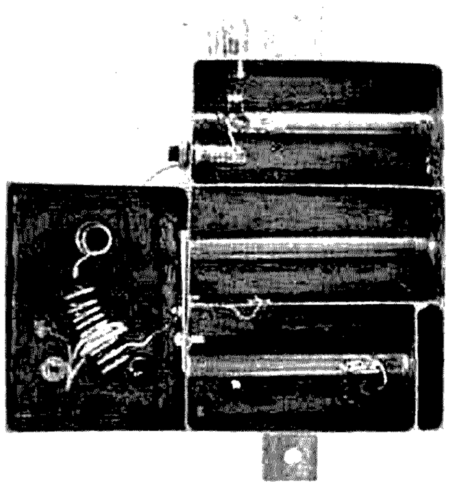
A small muffin fan was mounted on the rear door and the line voltage to it reduced with a series resistor. This fan helped take hot air out of the top of the cabinet, while the resistor kept fan noise down.

Details needing attention appeared after a period of operation. From the start, drive power requirements were higher than they should have been on any one of three tubes on hand, one supposedly "brand new." I made up a new set of cathode tank circuits of higher Q, but they improved matters only slightly. Finally, I obtained a fourth tube, and suddenly it became possible to drive to the legal limit with about sixty watts input. Moral: Make sure of your linear tubes!

It would have been nice if the second band-switch position for increased frequency coverage in the cathode circuit could have been used to shift the plate tank to a new frequency range. Thus, instead of requiring retuning for a shift from cw to phone on 3.5 and 7 MHz, the move could be accomplished as it is on 28 MHz.

I will be happy to correspond with anyone wanting to discuss any of the ideas incorporated in this linear amplifier. The courtesy of an sase would be appreciated. Let's keep the on-the-air testing to a minimum; no use dumping our garbage on somebody else's lawn!

ham radio



solid-state 2304-MHz converter

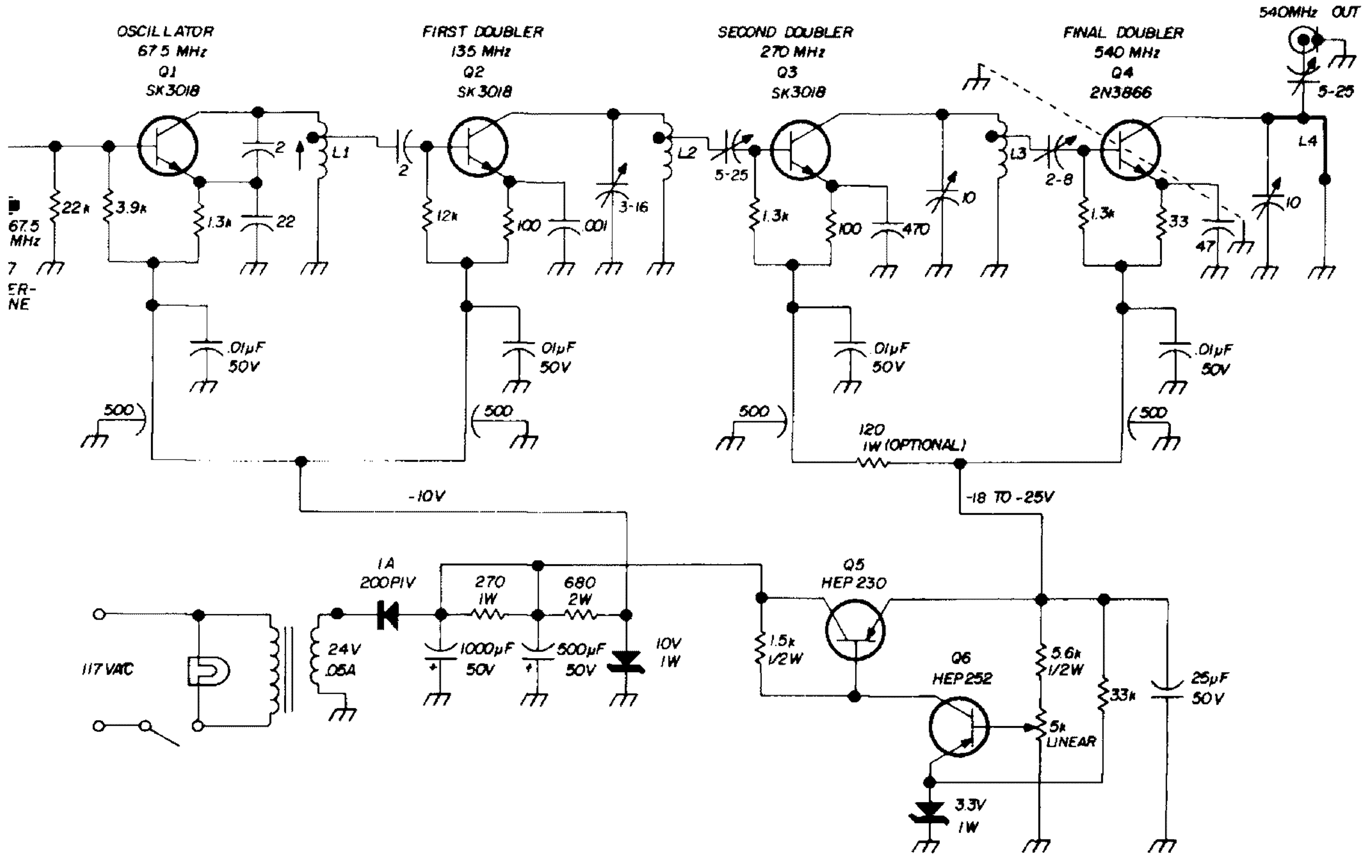
This high performance converter for the amateur 2300-MHz band boasts a noise figure of 8.5 dB when used with a low-noise 144-MHz i-f strip

Walter Stanton, K2JNG, Douglas Moser, WA2LTM, Dolph Vilardi, WA2VTR

There have been several articles covering converters and preamplifiers for the 1296-MHz amateur band, so amateurs have had several designs from which to select equipment for this band. Construction articles for 2300 MHz have been very scarce in the American literature, and most have involved designs which are often beyond the ability of the amateur to reproduce. With this in mind, K2JNG set out to design a converter for this increasingly popular band. His design had to be simple to construct and at the same time give adequate performance. Modification of the classic 1296-MHz design of W6GGV was the answer.[1]

The converter is presented as a practical unit to enable the amateur to operate on this band with reasonable success. This unit outperforms any of the wide-band surplus equipment usually available. Measurements of the converter's noise figure using professional laboratory equipment showed a noise figure of 12.5 dB with a 6-dB two-meter i-f, and 9.5 dB with a 3-dB two-meter i-f. Since noise figures of 2 dB are easily achieved in modern two-meter converters, an improvement of another I dB over these figures is readily possible.

Problems in building converters for these frequencies in the past have centered around the difficulty in obtaining local oscillator injection at the high frequencies involved without massive and elaborate local oscillators. With the advent of new inexpensive transistors and efficient low-power varacters, it is possible to obtain the necessary injection with a minimum of equipment, overall size reduction and simple construction. This



L1	11 turns no. 22 on ¼" slug-tuned form
L2	4¾ turns no. 18, 3/8" long, ¼" ID
L3	2½ turns no. 18, ¼" long, ¼" ID
L4	shim brass or copper, 3/8" wide, 23/32" long; length is in addition to tabs for soldering to capacitor and ground plane

fig. 1. Local-oscillator chain for the 2300-MHz converter. 540-MHz output is quadrupled to 2160 MHz with MV1622 varactor mounted in a trough line (see fig. 4).

converter is a collaboration of the three authors; WA2LTM was principally responsible for the oscillator chain, K2JNG and WA2VTR designed and optimized the multiplier and mixer trough-line components.

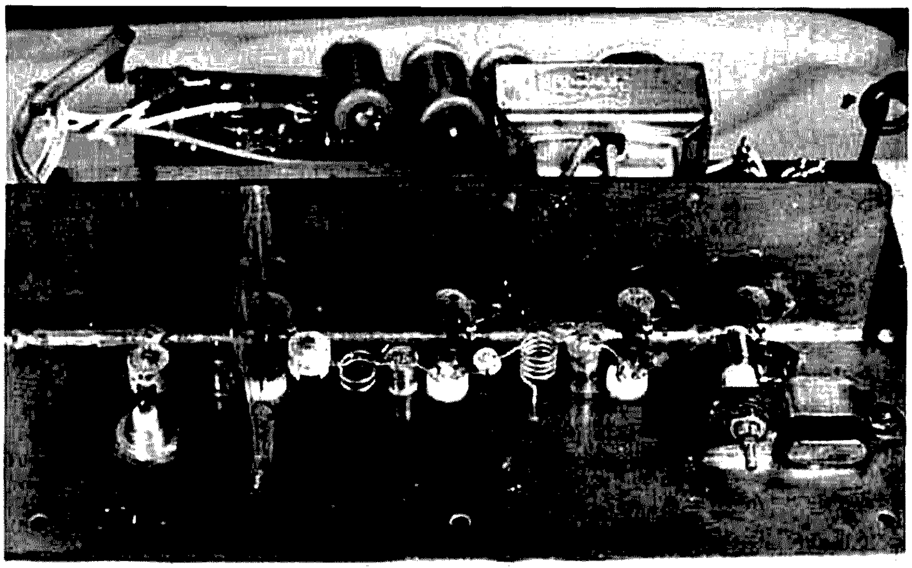
It should be pointed out that one of the major improvements was the HP2835 hot-carrier diode supplied by Hewlett Packard. This serves as a mixer and proved to be considerably superior to any other mixer diode tried at 2300 MHz. This diode is available for 90c from any Hewlett Packard sales office.

local oscillator

The local oscillator chain is transistorized for compactness. Although a variable regulated power supply is included in the description, it is not absolutely necessary. Any well-regulated supply capable of supplying 9 to 11 volts and 18 to 24 volts will adequately power the chain. This will provide approximately 250 milliwatts put at 540 MHz.

This 540-MHz signal is multiplied to 2160 MHz by the MV1622* epicap diode to provide sufficient injection at 2160 MHz to drive the hot-carrier diode mixer to 1 milliampere. Try any small varactor diode. Several versions of this converter have been built, and one of them uses a local oscillator injection frequency of 180 MHz which is multiplied in one step to 2160 MHz using a special abrupt-junction diode which is quite expensive and not easily available.

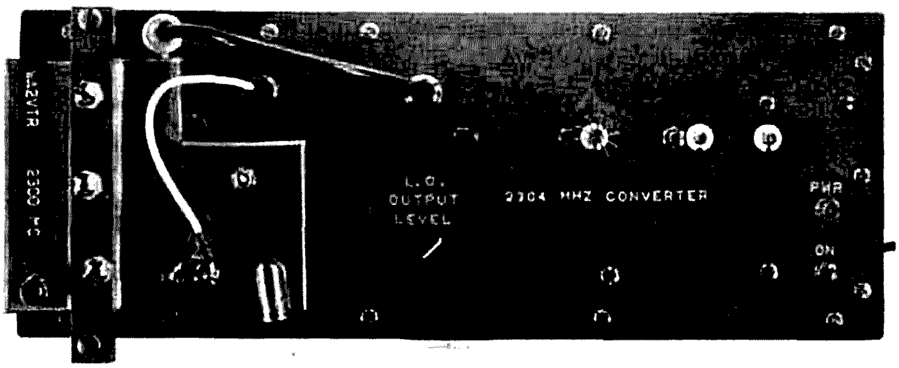
*Nearly any of the 1600 series diodes should work. The MV1622 is recommended. Some transistor junctions (base-emitter) will work as will any small abrupt-junction varactor rated at least to 8 GHz.



Construction of the 540-MHz local-oscillator chain.

Since the power required at 180 MHz is about 1 watt, many spurious signals are generated and this is not a recommended method of obtaining the required injection.

The local-oscillator chain as shown in fig. 1 uses six inexpensive transistors. The circuit is a conventional common-emitter configuration. Oscillator Q1 uses a 67.5 MHz overtone crystal and Q2, Q3 and Q4



Top view of the complete converter shows layout of local oscillator components.

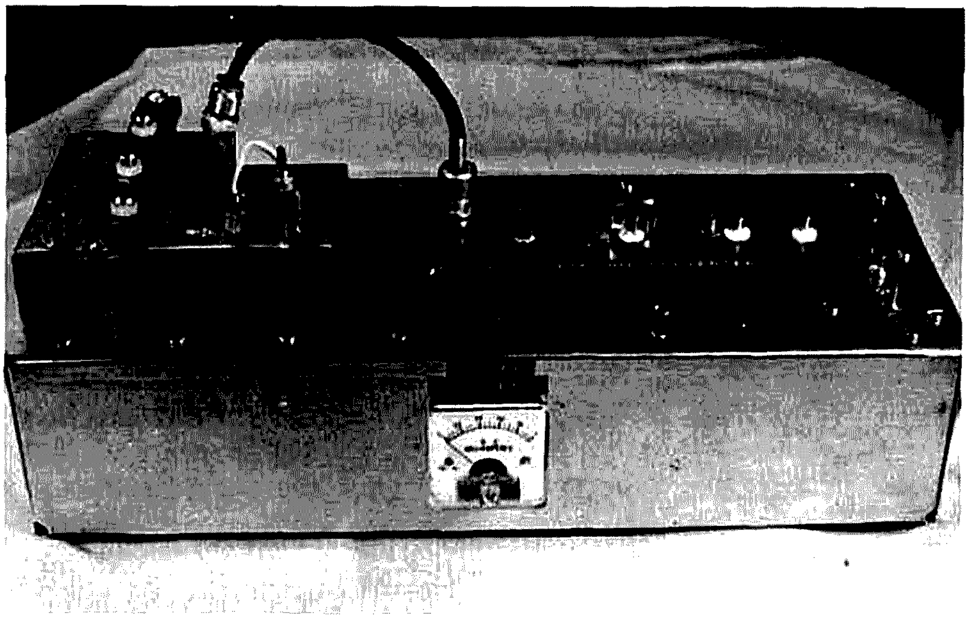
The photos and the schematic diagram of the oscillator chain should enable any experienced amateur to construct this portion without difficulty. It is recommended that the parts layout be followed closely, since it is important that parasitic and spurious signals be eliminated.

The trough line follows standard procedures and presents no major difficulties. For optimum rigidity use the heaviest brass sheet you can still bend. Flashing copper is not recommended, but good copper-clad material will work. Dimensions must be followed closely.

are frequency doublers, ending up with an output frequency of 540 MHz.

The supply voltage to Q1 and Q2 is regulated, while Q3 and Q4 have an adjustable supply voltage. The adjustable voltage allowed us to vary the output power of the local oscillator since it was not known how much mixer current would be obtained with the particular mixer diode used.

The circuit, as shown, can produce an approximate output power of 350 mW. If the variable coupling capacitors C1 and C2 are replaced by fixed values the



The complete 2304-MHz converter. Trough line is on top of chassis containing local-oscillator chain and meter for mixer current.

output may drop as much as 20% to 25%, but this should still be sufficient to drive the mixer diode to 800 μA.

It will be noted that coupling capacitor C3 has a low value for the frequency at which it is being used, but it was found that larger values made tuning more critical.

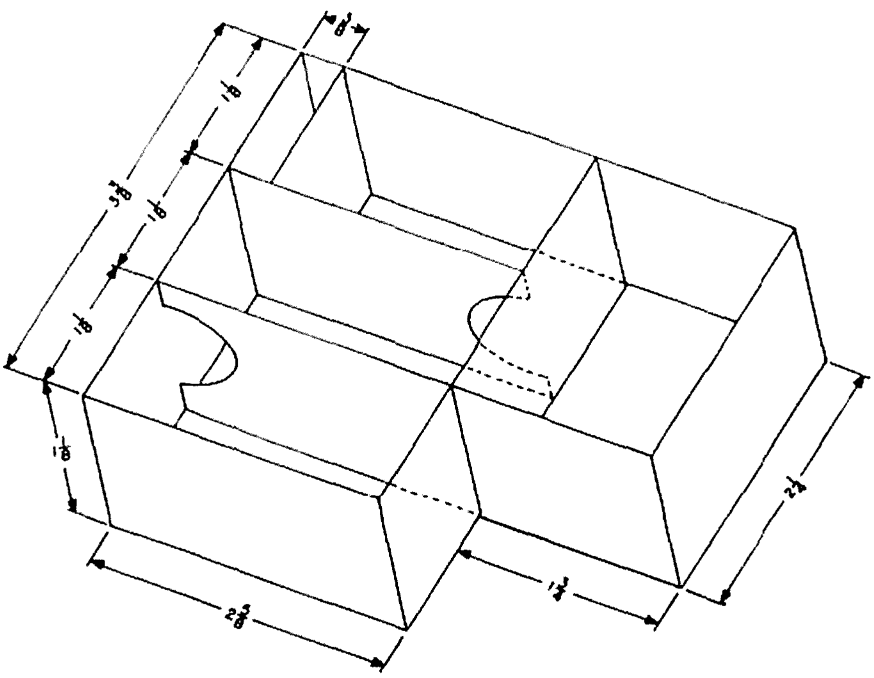


fig. 2. Internal dimensions of the 2304-MHz trough line. Material is sheet brass.

construction

Standard uhf building practices should be followed. Keep all leads as short as possible, especially the bypass capacitors. The local oscillator chain is built on double copper-clad board. Teflon sockets are used for all transistors to facilitate substitution and replacement. Shielding

was necessary only at Q4.

Stability was improved by mounting the crystal inside the chassis so air currents would not effect the temperature of the crystal. It was found that grounding the crystal case was important for good oscillator stability. Johanson ceramic capacitors are used in the parallel-resonant circuits of Q2, Q3 and Q4. However, any good glass or quartz piston capacitors should work as well.

Several types of transistors were tried in the circuit. A HEP709 (equivalent for the RCA SK-3018) was tried at Q1, Q2 and Q3 with poor results. However a HEP-75 was tried at Q4 and found to be to the dimensions shown. Hints on actual construction are presented in W6GGV's article or in the 1971 edition of the "ARRL Handbook" or "VHF Handbook" which describe the original 1296-MHz version. Techniques of soldering and construction are identical.

Since the HP2835 mixer diode is a pigtail version, the bypass capacitor for it differs from the original design (fig. 5). One lead of the mixer diode is bent to provide greater coupling into the signal trough. This is done experimentally after the converter is completed and aids in obtaining the best signal-to-noise ratio. Because the oscillator injection frequency

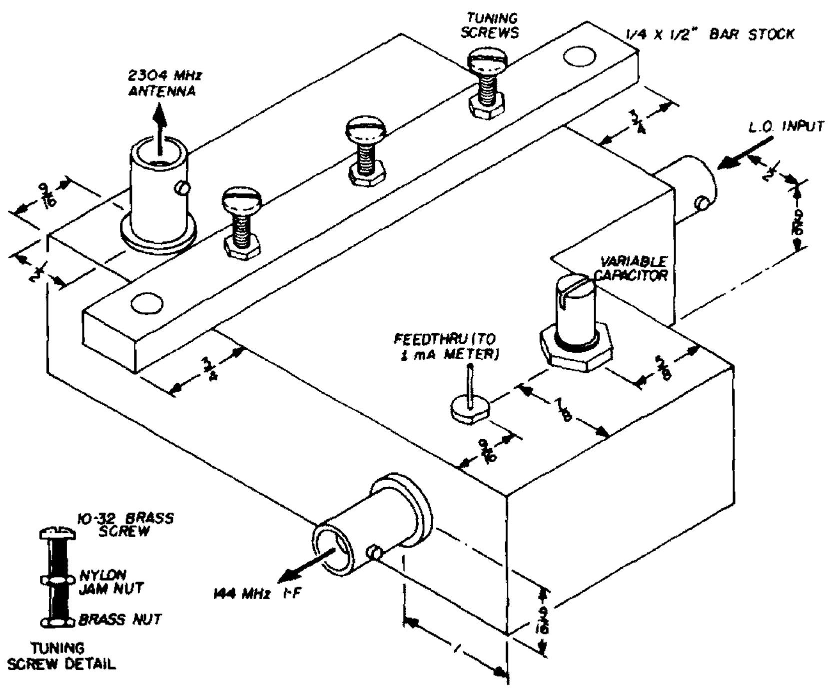


fig. 3. External construction details of the 2304-MHz trough-line converter. The ¼ x ½" bar stock is drilled and tapped for the 10-32 tuning screws used for C2, C3 and C4 (see fig. 4). Brass nut is soldered flush with the bottom of the tuning screw as shown in the detail. Nylon jam nuts hold the tuning screws in place after the converter is tuned up.

about equal in performance to the 2N3866. At Q5 and Q6, almost any inexpensive pnp transistors, with TO-3 and TO-5 cases respectively, may be substituted. This circuit has been duplicated three times by the authors and in each case it worked fine the first time it was tried.

the trough line

Brass should be cut and bent according is 540 MHz, the multiplier and filter troughs will *not tune to a* lower or a higher harmonic than the design frequency. Therefore, it is not possible to tune to any harmonic frequency other than the one desired (2160 MHz).

This simplifies tuneup, but in any case, it is best to start tuning with the capacitors screwed all the way in without touching the half-wave lines. It may be desirable to experiment with the antenna

input coupling, but the tap shown appeared to be as good as any other coupling which was tried.

Lap or otherwise carefully machine the bottom of the trough flat. It is important that the trough be mounted solidly on top of the chassis, making good contact at all points between the trough and the chassis. Failure to do this will result in difficult and erratic tuning; also, the trough line will pick up stray rf which causes variations in crystal mixer current.

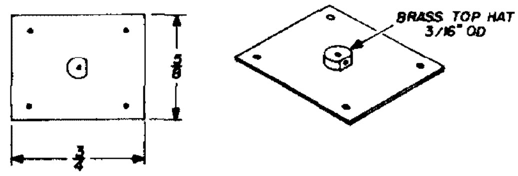
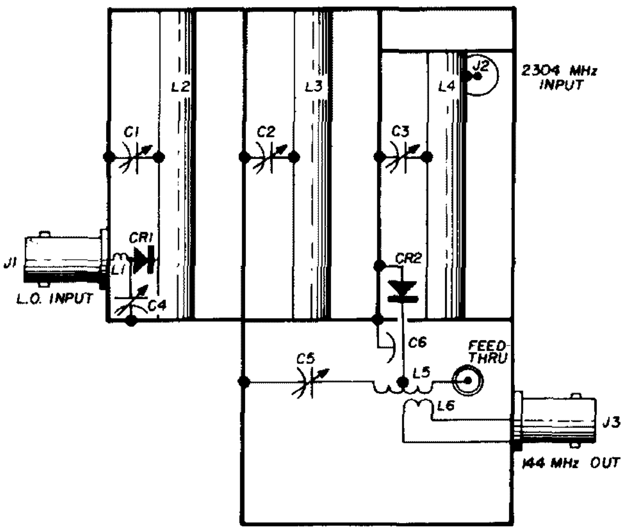


fig. 5. Construction of feedthrough capacitor C6 (see fig. 4). Brass plate is mounted on converter body with 4-40 nylon screws; use 0.005" to 0.01" Teflon between the capacitor plate and the converter chassis. The small brass top hat is drilled with a no. 60 drill and tapped on side for 8-80 set screw (for holding CR2). Alternately, CR2 may be soldered in place.

tuneup

Tuneup should proceed as follows. With approximately 200 to 250 milliwatts of 540-MHz signal injected into the trough line and a microammeter connected to the meter connection, tune the multiplier trough line for some meter indication. When this occurs, carefully tune the center trough (filter) for maximum indication. At this point, substitute a less sensitive meter, and tune both the multiplier and filter troughs for maximum indication; this should be somewhere in the vicinity of 1 milliampere. Peak up the L-network input to the varacter.

At this point, introduce a weak 2304 MHz signal into the signal trough. Connect the output of the two-meter i-f to a *good* two-meter converter which is in turn connected to the antenna input of a good communications receiver. Tune the signal trough for maximum signal. Tune the i-f tuning capacitor E for maximum S-meter reading. The input tap on the signal trough and the tap on the two-meter coil from the mixer may now be adjusted for best noise figure.



C1,C2,C3	10-32 brass tuning screws, see detail in fig. 3
C4	1-10 pF piston capacitor, miniature
C5	3-30 pF piston or air-variable capacitor
C6	bypass, see fig. 5
CR1	MC1622 multiplier diode
CR2	HP2835 mixer diode
L1	3 turns no. 30, 1/8" ID
L2,L3,L4	5/16" OD brass tubing
L5	10 turns no. 20, 3/8" ID, tapped for best noise figure
L6	2-turn link, position for best output

fig. 4. Construction of the 2304-MHz converter. Input signal from J2 is tapped 1/8 up from end of line L4. Feedthrough is 0.001 μF. Link coupling, L6, must be carefully adjusted for best noise figure; for more simple output arrangement, attach the cathode end of CR2 to the feedthrough end of L5, and use 0.001-μF capacitor from L5 to J3—tap capacitor down about 1 turn on L5. To tune 2287-MHz Appollo communications frequency, lengthen all trough lines by 1/16" and change local-oscillator frequency to 535.75 MHz (66.97-MHz crystal).

reference

1. H. M. Meyer, Jr., W5GGV, "A Crystal-Controlled 1296-Mc Converter," *QST*, September, 1962, page 11.

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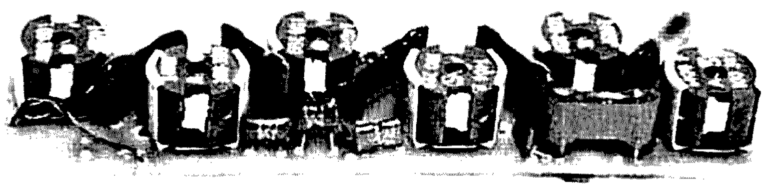


455-kHz filter

for amateur fm

Ferrite pot cores in a tailored filter design

G. K. Shubert, WAØJYK, 1308 Leeview Drive, Olathe, Kansas 66061

Fm receiving techniques have been the subject of much attention in the amateur literature.[1-5] While some of these circuits use high-frequency crystal filters, many are dual-conversion circuits in which selectivity is developed at lower frequencies.[6]

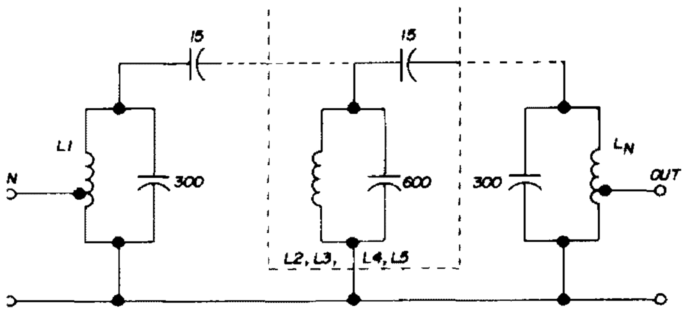
Crystal filters are commercially available, but LC filters are not. This article provides information for constructing a 455-kHz LC i-f filter with selectivity and impedance to meet fm receiver requirements.

Ferrite pot cores, although available for at least a decade, have not found widespread use in amateur radio circuits. An article by Hank Olson, W6GXN, describes pot cores and other ferrite and powdered-iron elements and how to use them.[7] The compactness and adjustability of ferrite pot cores, as well as their Q, put them ahead of the competition for fm receiver i-f strip application.

filter circuit

The basic filter is shown in **fig. 1**. Note that it has two LC end sections and up to four LC center sections. Less filtering is required by some detectors; others may require more, so the selectivity may be chosen by adding or deleting center sections.

The graph of **fig. 2**, which was plotted from a computer printout, allows you to select the number of LC sections, or poles, required for the desired selectivity.



L1, L_N	60 turns 9/41 or 18/44 Litz wire on ferrite pot core
L2	40 turns 15/41 or 30/44 Litz wire on ferrite pot core

See text pot cores suitable for this filter.

fig. 1. Filter circuit for fm receiver i-f strips using LC elements. Coils are wound on gapped ferrite (pot) cores. Theoretically, any number of center sections may be added to increase selectivity at the expense of increased insertion loss.

Eight LC pairs seem to be the maximum for amateur use.

The filter is designed for the popular amateur "bellyband," i. e., i-f circuits for ±7.5 kHz deviation. However, the circuit selectivity may be increased or decreased by increasing or decreasing the value of the 15-pF coupling capacitors in **fig. 1**. Adding several center sections will tend to narrow the passband slightly. This may be compensated by increasing the value of the coupling capacitors.*

Input and output impedances also may be tailored by choosing taps from **table 1**. These impedances need not be the same. The filter is an ideal circuit with which to transform impedances.

construction

The filter elements should be laid out linearly to avoid stray coupling. Pot cores in clamps are inherently well-shielded. Coupling between cores is not a serious problem, even when closely spaced. The pot-core clamps should be connected to the common side of the circuit.

tuning

Tuning the filter is simple if this procedure is followed. Note that tuning for maximum output *does not* yield the optimum filter shape factor. The method outlined was developed by M. Dischal, a pioneer in electric wave filter theory.

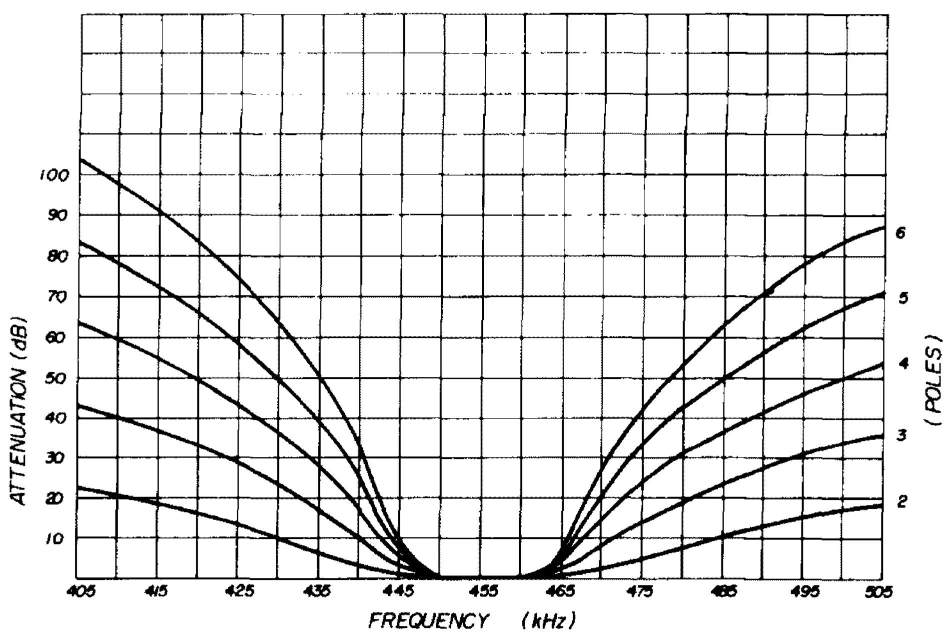


fig. 2. Selectivity curve for choosing number of filter poles for desired selectivity. Data were plotted from computer printout.

*Filter insertion loss may be of concern to some readers. Author's computer printout shows filter insertion losses as follows: 2 coils, 0.77 dB; 3 coils, 1.5 dB; 4 coils, 2.27 dB; 5 coils, 3 dB; and 6 coils, 3.77 dB. *editor.*

The equipment required need not be expensive nor exotic. A 455-kHz signal is required. Since the signal need not be tuned or swept, a crystal oscillator is ideal (perhaps a Motorola test set). The signal-source impedance should be similar to the filter input impedance, but it is not critical. If the signal generator is a low-impedance device a series resistor may be added. If the signal generator is a high impedance unit, then a shunt resistor will be needed. The meter must be capable of reading the output of the signal source with about 30 dB additional loss. It's possible to use an existing i-f strip with an S-meter. A sensitive dc meter may be used with a crystal diode as a detector. The meter is used only as a peak and null detector.

table 1. Impedance vs tap turns.

impedance (ohms)	no. turns
25k	no tap
10k	38
5k	27
2k	17
1k	12
500	8
50	2

Ferroxcube, part no. 1408 CA 100 3B7.
Magnetics Inc., part no. 1408 AL100 'D' material.
Indiana General, part no. TC7-01.

Be sure to specify an adjustable pot-core assembly with the least expensive clamp and a single-section bobbin.

This filter may also be constructed more economically, but with slightly

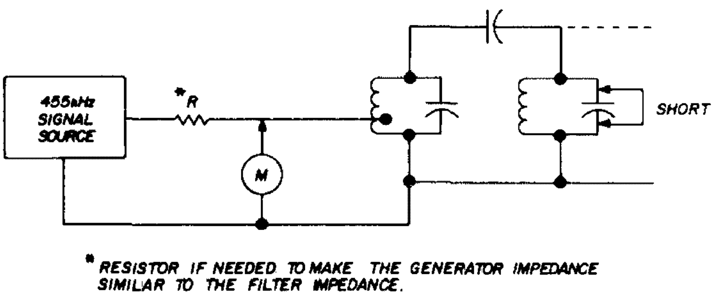


fig. 3. Instrumentation for tuning filter. Resistor R may be required to match signal generator impedance to that of filter, as explained in text.

Connect the signal source, filter, and meter as shown in fig. 3. Adjust the frequency to 455 kHz and forget it. Short across the second coil and tune the first coil for a peak. Short across the third coil and tune the second coil for a null. Continue to move the shorting clip down the line, and tune the immediately preceding coil to peaks and nulls alternately. The last coil is tuned with no shorting clip.

Should there be difficulty in tuning a coil, remember that the above equipment setup may be used to tune a single coil by peaking the coil. If a variable oscillator is used, it may be tuned to find the circuit resonant frequency. A grid-dip meter is useless with pot cores.

The ferrite pot cores are available from several sources. Each of the following cores was tried with similar results:
Ferroxcube, part no. 1408 CA 100 3B9.

reduced performance and much more difficulty in tuning, by using powdered-iron toroids. The T80-3 core would be a reasonable choice. Variable trimming capacitors may be used for tuning.

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ham radio

improved two-meter preamplifier

An easy-to-build non-neutralized two-meter preamp

Jerry Vogt, WA2GCF, 182 Belmont Road, Rochester, New York 14612

As any writer knows (and I am a technical writer by trade), the hardest part of writing anything is just getting started. I've been meaning to write an article on the subject of two-meter preamps for quite some time, and I've never been able to think of a good gimmick to start it off with. I like to think that communications, to be effective, has to be somewhat like one person talking to another, with nothing holding you back. The constraints of writing military technical manuals prevents much of that type of thing, but I think that a ham journal can be an exception. So here is a little bit of my thoughts on the subject of preamps. Of course, everything you ever read has to be taken with more or less of a certain seasoning ingredient, so here is some background.

Not being a design engineer, I make no great claims as to scientific discoveries on preamps. Personally, I don't believe in a lot of the magic I read about; however, there are good and bad points which can be weeded out from various circuits which come along. Having been involved in a club project to build preamps, I have experimented with several circuits over the last few years; and I came to a surprising conclusion, which will be the subject of this article.

brief history

To review some recent innovations, and some of the circuits of the past as well, I should start with the frame-grid tube. That was the first new thing to come along in quite a while, for receiver front ends, when it was introduced several years ago. It boasted high Gm and low noise figure. Of course, there was the old 417A which did a pretty good job at that time, but you had to scrounge them from your friendly telephone company because you couldn't afford the price of a new one. I guess that I still didn't know for sure what Gm and low noise figure meant, but everyone said they were good to have; and we still hear it today. Except, now we take those features for granted.

Then, along came the nuvistor. It supposedly had the same features, only more so. But then, it had the drawback that all tubes have, namely that it takes extra wiring and power, that it has to warm up, and that it will wear out. It did make a pretty good front end, however. Then came bipolar transistors.

The appeal of getting away from tubes was very strong, and everyone wanted to make a completely solid-state receiver. Like a lot of new innovations, it was a long time growing up. Unfortunately, quite a few terrible front ends were built just for the sake of using transistors. The biggest problems were intolerable overload characteristics and poor noise figure. Several years passed and we finally got the field-effect transistor. This solved the overload bug, and before long, the 2N4416 was introduced, which gave us good noise figures.

So that was it for awhile; but alas – what had we learned from all the work that our predecessors had done to make an easy-to-duplicate circuit with good operating characteristics? Not a darned thing! Article after article came out with single, neutralized field-effect transistor amplifiers in converters and preamps. Baloney! I think back on the many occasions during which I had built converters and preamps and practically tore my hair out trying to neutralize the darned things. I don't have a single good thing to say for them. They were inexpensive and they were simple to build if you could figure out what value of inductance should cancel the output-to-input capacitance, but they certainly didn't tune up well. And afterward, if you changed the line voltage the feedback capacitance would change also. Change devices after a failure. All your values must be changed to make it play again. I built 25 preamps for our club with such a circuit. I think I aged ten years. Never again.

Then, along came the dual-gate mosfet. That is another subject. I like to call them moose-fets. I guess I kind of feel sort of cold about a transistor that blows before I even get it in the circuit. I heard so many tales about mosfets blowing that I've never tried them. So I won't criticize their performance. They probably work very well once you solder their delicate little legs to your board, but you couldn't prove it by me.

There are probably many readers who agree that one branch of Murphy's law says that even on a day with 101% relative humidity, you will have sufficient static electricity to blow every mosfet you even think about touching. Of course, there are proper methods to prevent said disaster from ever occurring; but that's kind of like picking up broken glass. There is no excuse for being cut, but you still get nervous. So much for slandering "the best thing yet in semiconductors." What can we learn from all of these things?

design criteria

I set out to design a good compromise preamp. One that I could mass-produce without getting ulcers, and one that I could recommend for others to build. So I sat down to find out what I wanted. I didn't want tubes, for obvious reasons, but they had worked well by past standards. I didn't want neutralized junction fets – at least, not any more! I didn't want ticklish little fellows which blew

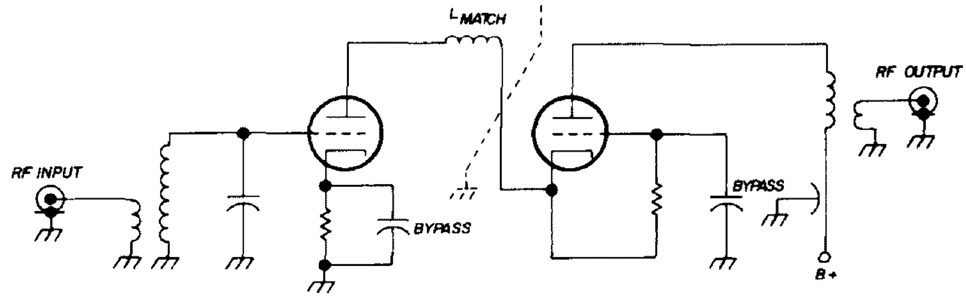


fig. 1. Classic vacuum-tube cascode amplifier for vhf receivers.

before I got them in the circuit; and I was too proud to think that I was going to get sucked into playing little games to keep the little devils from blowing up to defy me. So I referred to my library, which has many old books with not-too-well hidden secrets. Of course! Why didn't I think of it before?

the jfet

I set out to design a cascode preamp with jfets. A natural! Jfets are fine devices, readily available, and come in many flavors. Use a cascode circuit like in the days of tubes (you remember what those were) and you don't have to neutralize. Jfets have all of the required characteristics: good gain and low noise figure to hear with, and good square-law characteristics so you don't hear the taxi service or fm broadcast station two blocks away. Some educated types even admit now that it's possible that mosfets are only better for preamps because they are inherently cascode. Is that all? We can do that ourselves. Of course, if you want agc, then dual-gate mosfets are the thing. But, who needs agc? Ever looked through designs for the new ham transceivers and the commercial transceivers for business band? They don't use agc; not as a rule.

design hints

I am a great circuit snatcher. I never build anything exactly the way it was done in an article or in a piece of commercial gear. But I snatch a lot of ideas. I combine a stage from some past article with a few circuits from commercial radios. And I do a lot of homebrew work. At least, that's what my wife said the last time I saw her. Now, I want to let you in on something ironic. One day at the plant I was belly-aching to one of our better design engineers about all the grief in duplicating the ham-type circuits. The commercial ones are usually easier to copy. I told him exactly what I thought about neutralizing and combining functions in one stage (like oscillators and multipliers). He made me stop and think. He matter-of-factly designs things like a solid-state am-fm *kilowatt* 150-MHz power amplifier for a living. He told me that the biggest mistake that hams make in selecting circuits is that they count parts. They don't count labor or repeatability, just parts. This is probably due to the traditional pocket-book problem. But be practical! I have been using some of his ideas awhile, and I want to tell you that the pleasure is back in homebrewing for me now.

The philosophy is based on two premises. First, when designing a circuit, use more parts if that's what it takes to make it easier to build. Consider that your time is valuable. Don't waste it trying to figure out how to save five cents on a resistor. Even if it costs you an extra buck for the second transistor, it sure outweighs wasting a whole afternoon trying to make a silk purse out of a sow's ear. Secondly, transistors don't have all that much gain that neutralizing should be necessary. Mismatch a little. Throw away that last ounce of gain. Use more stages or devices instead. You'll be a better designer for it.

In industry, it is mandatory that you don't depend on every bit of gain. You have to consider other characteristics and repeatability. My friend's basic philo-

sophy is, "If it oscillates, swamp it." In other words, if you have so much gain that it takes off, load it down with resistors until it stops. Then, if you need more gain, add another stage. If you need more selectivity, use a multi-pole filter or a different type of filter.

cascoded jfets

Having proven to myself that these premises do hold true, I took exactly that approach in designing the cascode jfet preamp. I didn't worry too much about noise figure and overload resistance. These are difficult to measure quantitatively. They pretty much take care of themselves if you use the traditional rules and use good devices. The biggest problem was how to make a cascode circuit.

For those who have forgetten what the old tube-type cascode circuit looked like, refer to fig. 1. I won't bother to explain all the details of circuit design; you can get that out of any old handbook. Basically, one stage acts as a voltage amplifier, and one acts as a current amplifier. The first tube's plate circuit is loaded down heavily by the cathode of the second. Therefore, the unneutralized common-cathode first stage doesn't oscillate. The grounded-grid second stage, of course, doesn't oscillate either. The two act together to give you about the same overall gain that a neutralized single-stage amplifier would yield.

Trying to visualize what the equivalent circuit in field-effect technology would be brings up an interesting point. Tubes of the type used in such circuits were usually of the close-spaced element design, apparently to obtain the high Gm desired. Therefore, it was necessary to use about half the normal B+ voltage of conventional tubes. That was great, though, because two tubes should be connected in series anyway to make a cascode circuit; so they were also connected in series for dc operation.

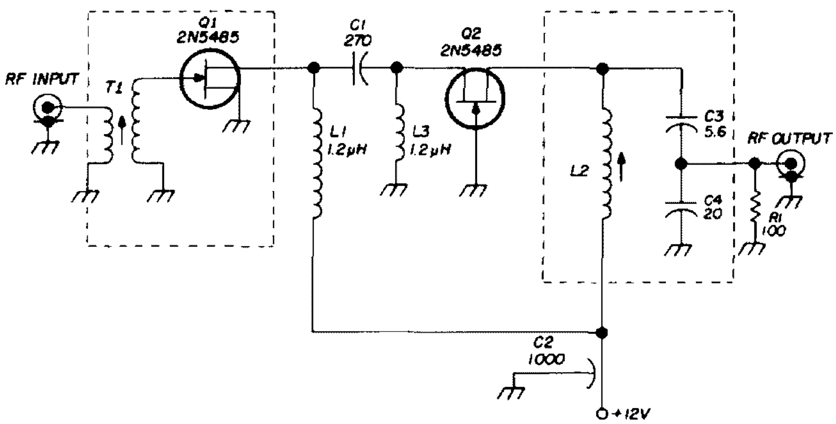


fig. 2. Schematic of the improved jfet preamp. Inductance values are discussed in the text.

Not so with fets. They work just fine on 10 to 15 volts, so why use a dc-series configuration? Indeed, the problems of determining the correct biasing method are also complex. The solution was to use an ac-cascode circuit. Further simplification was discovered that made biasing even easier. Selection of the 2N5485 resulted in obtaining optimum operation with no bias at all. No series source resistor is required, and therefore, no bypass is required in the source circuit.

actual circuit details

So much for my editorializing! The finished design is shown in fig. 2. The preamp is an ac-coupled cascode jfet amplifier. It provides 16- to 25-dB of actual gain between terminals. Empirical results (meaning that about 50 of them are already out in the field) seem to indicate that they work well. They improve reception on virtually all receivers.

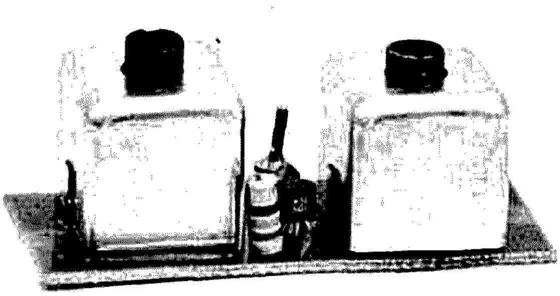
The older receivers, of course, use tubes; and good low-noise gain to swamp out the noise in the present front end of the receiver is bound to make an improvement.

Oddly enough, even the latest solid-state equipment on the market today can be improved, simply because the manufacturers either cheapen the design to be competitive or, in the case of commercial radios, the manufacturer also has to meet a cross-modulation spec. In the case of the latter, radios are sold for use under unknown rf-pollution conditions, so the manufacturer is watching out for selectivity too, and is willing to sacrifice a few microvolts of sensitivity. In most cases, the commercial receiver will listen to a quarter kilowatt up on a prime antenna location, so who cares.

Of course, you don't get anything free. You do sacrifice some overload resistance by driving the front end harder with a preamp. In most cases, though, the somewhat increased overload susceptibility is negligible, especially in mobile sets.

Refer to **fig. 2.** The input of the preamp uses transformer coupling, with the secondary tuned by the 5-pF input capacitance of the fet. Coupling is such that a 50-ohm source resistance loads the gate of the fet to approximately 1000 ohms, a typical design parameter. To provide ac coupling to the next stage, the drain of the first stage is shunt fed through L1, a 1.2-μH choke. The signal is coupled to the source of the second stage through a bypass-type capacitor (C1), and the source of the second transistor is tied to ground for dc through L3.

As previously mentioned, the second transistor, being fed at the source, loads down the drain impedance at the first transistor. The second stage runs as a grounded-gate amplifier, and its drain load impedance is established by a capacitive divider across the output coil, which is tuned by a combination of the divider (C3-C4) and the 2-pF output capacitance of the device. The capacitance ratio is set up to provide a load impedance at the drain of approximately 5000 ohms, another typical design parameter for such preamps. The B+ input to the output coil is fed through a bypass capacitor consisting of a solder-in feedthrough capacitor. This also provides a convenient tie point for connection of the B+ wire.



View of the completed preamp with coil shields in place. L2 is at left, L3 in the center (looks like a resistor) and next to that is Q2 and T1. Directly behind Q2 is C1 and C2.

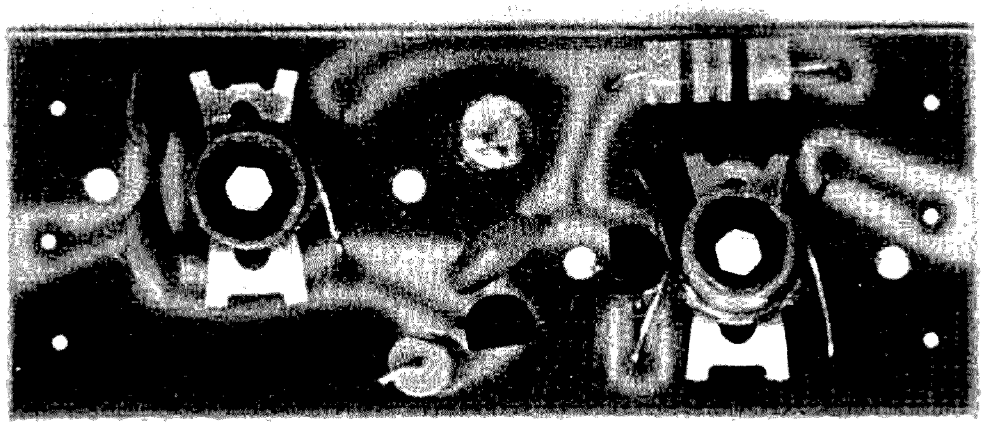
A feature of this design is the simple tuning after construction. Because no neutralization is necessary, and because the input and output coils are completely isolated, tuneup is a simple matter of peaking the coils. Of course, that depends on the proper range of the coil to begin with. In the model illustrated, the coils consist of four turns of number-22 Solder-eze wire, each with a one-turn input link on the input coil. Coil forms consist of paper-phenolic impregnated with silicon wax. Tuning slugs are of the ¼-20 type, 3/8-inch long, with an internal hex thru-slot to accommodate a standard J-tran tool. Slug material is iron-8. Special Tinnerman coil retainers are used to hold the forms in the printed circuit board, and special ¾-inch coil shields are used to cover the coils and some of the other parts as shown in the photo. Don't worry about the special parts, however, since I

have arranged to make them available.*

installation

Each end of the board has two holes provided in the ground plane. One hole at each end is used for mounting, and the other is used as a ground connection for the coax input or output cable. Thus, the unit can either be mounted on standoffs on opposite corners, or it can be mounted with L-brackets attached to two holes on the same side. The latter method works out well in the Motorola tube sets which have rectifier cages. The preamp can be hung on its side from two brackets installed at the top of the cage which is located near the receiver input connector.

As you may be aware, the stripped and separated ends of a piece of coax have inductance, as does any straight wire. Keep the stripped ends extremely short to avoid loss and pickup of noise and stray rf. A good method of connection for the braid is to wrap the exposed braid end with number-22 bare wire, right around the cable. Tie or solder the ends to secure them around the cable, and then solder the bare wire to the board.



Components side of the printed-circuit board with coil shields removed. L1 looks like a resistor in the upper right corner; C4 is above C3 all the way on the left. The board, with an overall length of only 2½ inches, is very straightforward.

On the subject of power, any source of filtered 10 to 15 volts can be used. Gain is relatively flat above 10 volts, so regulation is unnecessary. Also, tuning is not particularly affected with changes in voltage as is usually the problem with the neutralized types. Power connection is made to the top terminal of the feedthrough capacitor. (Note that the cold end of the winding on L2 is soldered to the board and continued on to be soldered to the bottom terminal of the feedthrough capacitor.)

A word of caution is in order. One of the previous builders, figuring the current drawn by the preamp as being about 5 mA (which it is), decided that he could use a dropping resistor to reduce the receiver's 200-volt supply to 12 volts. His calculations were fine, but he should have

*The following are being made available in conjunction with this project. A complete parts kit, including the G10 pc board already drilled, is available for $6.00 postpaid. Completely built preamps, tuned to any frequency in the 144-172 MHz band, are available at $10 postpaid. Quantity prices are available to clubs to allow clubs to make a profit. Factory built preamps can be returned for repair (prepaid) for a fixed repair charge of $3, anytime during the first 90 days. Contact HAMTRONICS, 182 Belmont Road, Rochester, New York 14612.

used a zener diode in addition to the resistor. The transistors failed, of course, due to high voltage surges. Another word of caution, which should be obvious, is that you can't transmit through your preamp. Remember that when you outboard a preamp on the back of your new Japanese transceiver!

tuning

After testing your preamp on a signal generator and a 50-ohm load, you should repeak the coils slightly when the unit is installed in the set. At two-meters, a slight amount of reactance in your external circuit will detune the preamp slightly. But, being a cascode design, retuning is very simple; you need not worry about oscillation. (You may notice that no attempt was made to keep you from tuning it; such is not the case with a popular preamp which is soldered shut at the factory to prevent the neutralized circuit from taking off when you play with it.)

The cascode circuit shouldn't oscillate. However, remember that any circuit exposed to another circuit at the same frequency may cause the combination to oscillate. Some of the inexpensive radios on the market today (and I don't know how they do it) have absolutely no shields around coils in the frontend. This may be fine if you don't have a lot of gain (which may be the case); but if you add a preamp anywhere near the frontend coils in such a unit, I'm afraid that you will have to provide a shield box around the preamp to prevent pickup from the receiver's coils. Otherwise, you may find that the whole set takes off when you upset the apple cart by adding your preamp.

Well, that's the story of what you go through to design a preamp. Whether you build one or not, I hope that some of the hints and kinks will help you in building your next transistor rf project. If you have questions, feel free to write to me. Please enclose a sase, and I will try to jot down a few notes to help you if I can.

ham radio

reciprocating detector

This novel circuit has many advantages over conventional detectors since it automatically adjusts its bfo level in proportion to the average signal level

Stirling Olberg, W1SNN, 19 Loretta Road, Waltham, Massachusetts 02154

A paper presented at the International Communcations Conference[1] during June 1971 described a synchronous detector which should be of interest to amateur radio operators. The circuit was designed by R. S. Badessa while working on a project at Massachusetts Institute of Technology. Mr. Badessa made further investigations through support given by Damon Corporation where we both are employed.

The circuit was primarily designed for double-sideband, suppressed-carrier (dssc) detection. It has been used in several communication receivers as a second detector and exhibits features which make a superb demodulator for CW, a-m, dsb and ssb.

The name, "reciprocating detector," seemed appropriate to the inventor who described the detector's operation thus: "Because a suppressed carrier wave assumes either of two diametrically opposite phases in sequence, the detector channels these into a smoothly rotating reference vector."

The design features a carrier-synthesized reference signal and therefore does not require an external beat-frequency oscillator. Because of other characteristics of the circuit, impulse noises are rejected. Also, the average reference level is proportional to the average signal level. The CW DX chaser and moonbounce enthusiast can appreciate the desirability of this important feature of the detector when he remembers what bfo hiss noise does to a weak signal as it goes into a fade.

circuit operation

To aid in a brief discussion of the circuit operation refer to **fig. 1**, a signal flow diagram. An rf signal from the receiver i-f system is presented to a half-wave diode detector. The detector provides a signal current source which is fed to an electronic bidirectional switch. The two outputs of this switch are directed into the inputs of a differential amplifier. The amplified output is fed into a narrowband i-f filter and a low-pass filter. The narrow-band filter, approximately 500-Hz wide, is coupled to a phase splitter which returns the outputs to the inputs of the bidirectional switch; this filtered signal is the reference. The

low-pass filter allows the audio component to pass into the receiver audio system.

Previous experiments allowed the investigator to choose, by means of a selector switch, existing detectors in a Collins 51S1 receiver or the reciprocating detector. Later, simultaneous records were made from each detector for comparison.

The circuit diagram, **fig. 2**, is for incorporation into a Drake R4A receiver. It is possible to use the same circuit in any other communication receiver with appropriate modifications to the filter FL1.

The modification to the Drake R4A *involves rewiring the* CW/SSB/AM selector switch. All changes are temporary; this allows the receiver to be restored to its original state. At W1SNN the detector was permanently installed in the R4A. The crystal switch, designated S4A/B in the Drake Manual, located on the left side of the R4A receiver, was disconnected from the vfo/crystal circuitry. The vfo was permanently connected, freeing the switch used for S1. A dpdt toggle switch can be externally mounted in a convenient location on the operating table if you don't want to use S4A/B for this modification.

The schematic diagram shows the rewiring of the CW/SSB/SW switch. It should be wired exactly as shown. Otherwise, problems with the receiver bfo will result. The bfo must be off when the reciprocating detector is switched in or a steady beat will be heard due to bfo leakage into the receiver i-f circuits. This switch is designated S2 rear in the schematic diagram and in the Drake instruction manual.

The power supply circuit shown in fig. 3 provides the required voltages. The voltages are higher than called for in the diagram. It is important that the voltages be very nearly the same level; in excess of six volts is permissible, provided that the two are equal. They should not exceed 10 volts, however.

Resistor R24 must be included to complete the voltage drop through the voltage divider when the bfo is removed. Therefore, when the reciprocating detector is switched in the connections to S1 must be made as shown.

The direction in which the windings for FL1 must be wound is important. If they are in the wrong "sense" the filter will not operate.

tuneup

FL1 has a small adjusting slug which tunes the filter to the center frequency, 50 kHz. Put the CW/SSB/AM switch on ssb and S1 on the receiver's own detector; tune in an ssb station, and switch S1 to the reciprocating detector position. If the voice sounds higher or lower in pitch than

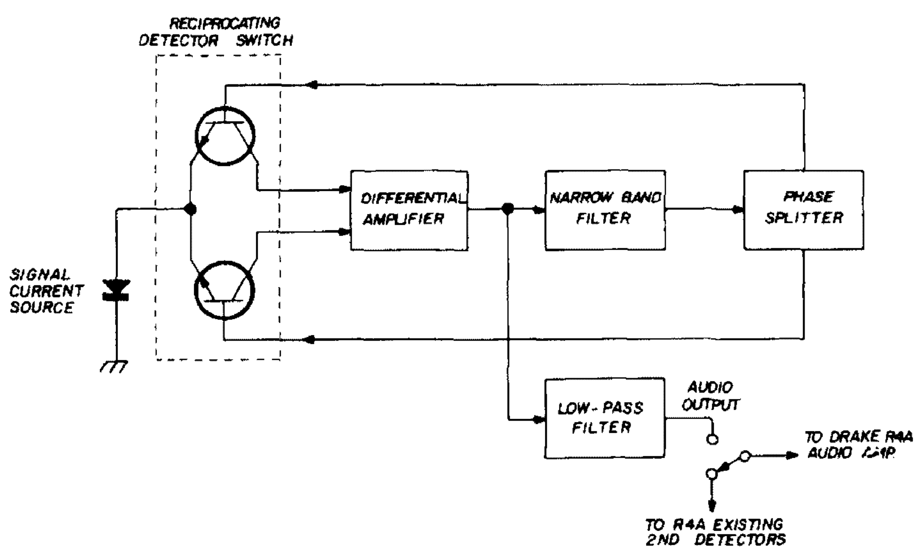
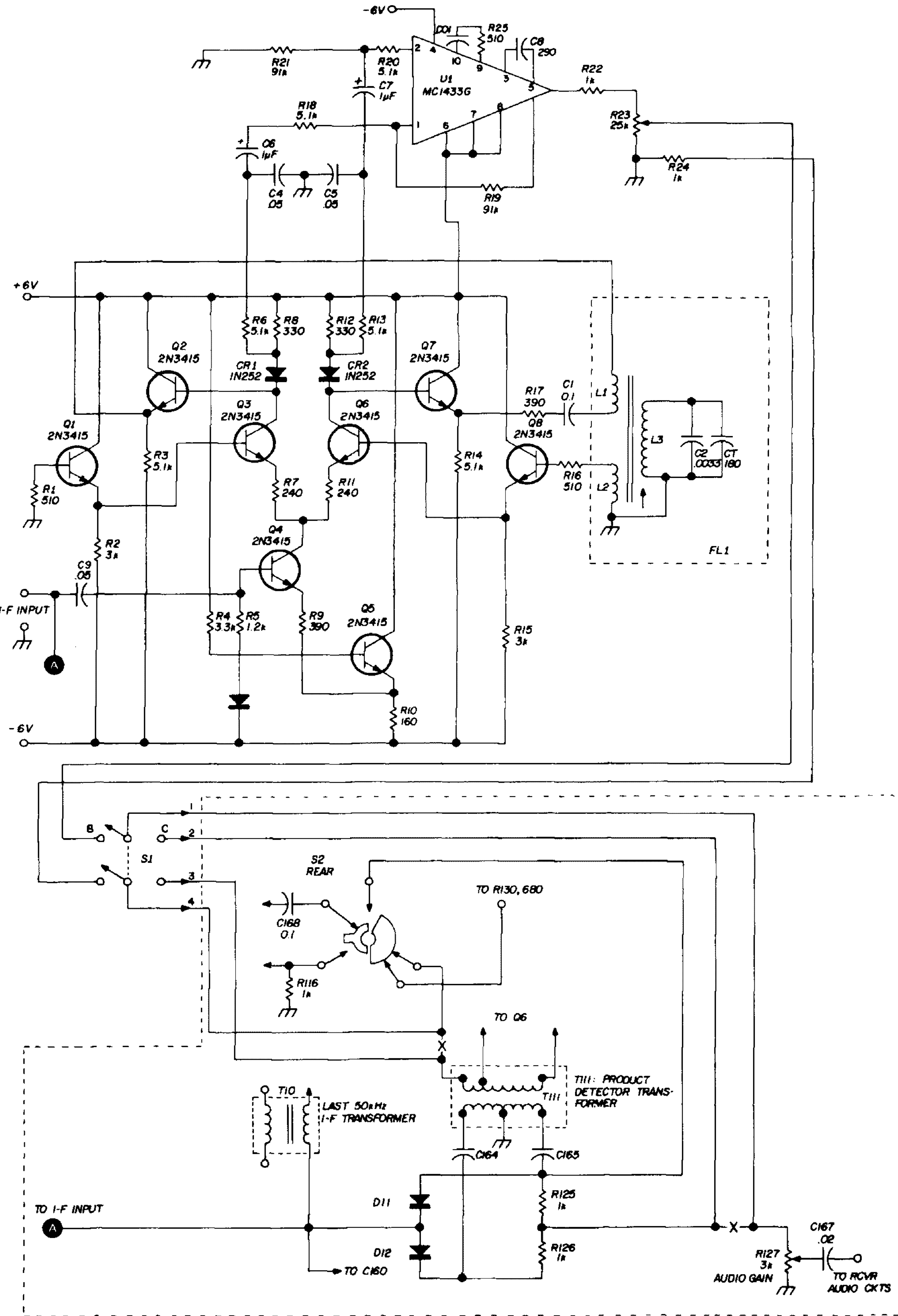


fig. 1. Block diagram of the reciprocating detector.


-6V
R21
91k
R20
5.1k
C01
R25
510
C8
290
U1
MC1433G
C7
1µF
R18
5.1k
C6
1µF
C4
.05
C5
.05
R19
91k
R22
1k
R23
25k
R24
1k
+6V
R6
5.1k
R8
330
R12
330
R13
5.1k
Q2
2N3415
CR1
1N252
CR2
1N252
Q7
2N3415
Q3
2N3415
Q6
2N3415
R17
390
C1
0.1
L1
Q8
2N3415
L3
C2
.0033
CT
180
Q1
2N3415
R1
510
R3
5.1k
R7
240
R11
240
R14
5.1k
R16
510
L2
FL1
Q4
2N3415
R2
3k
C9
.05
I-F INPUT
R4
3.3k
R5
1.2k
R9
390
Q5
2N3415
R15
3k
A
R10
160
-6V
B
C
S1
S2
REAR
TO R130, 680
C168
0.1
R116
1k
TO Q6
T10
LAST 50kHz
I-F TRANSFORMER
T111: PRODUCT
DETECTOR TRANS-
FORMER
T111
C164
C165
TO I-F INPUT
D11
R125
1k
D12
R126
1k
TO C160
C167
.02
R127
3k
AUDIO GAIN
TO RCVR
AUDIO CKTS

normal, adjust until the voice pitch sounds correct, by switching back and forth between the two detectors. Adjustment is complete when no difference in voice pitch is noticed. R23, an audio gain trim potentiometer, can be adjusted at the same time, rocking the switch in the same manner; this pot will set the audio output level of the reciprocating detector the same as the receiver detector output level.

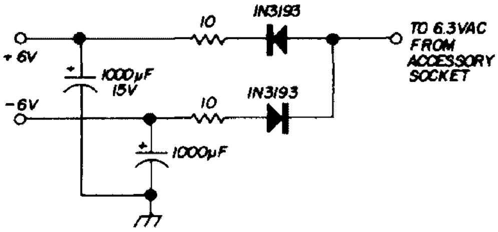


fig. 3. Simple power supply for the reciprocating detector.

operation

When the detector installation adjustment is complete, you can compare signals by simply flipping switches. At first very little difference will be noted in the comparison between detectors.

On 160 meters, with the receiver a-m detector switched in and the noise blanker off, an a-m signal was tuned in. Some interference from a Loran station was present. Switching to the reciprocating detector a beat signal was heard; re-tuning the signal very slightly produced a zero beat which has a very narrow lock-in range. No difference in audio quality could be noticed, and the Loran signal was greatly subdued.

Switching in the noise blanker eliminated the Loran signal pulses completely; switching back to the receiver a-m detector with the noise blanker on, the pulses were subdued but very difficult to copy through. The reason the reciprocating detector eliminated the Loran pulses is because the reference filter Q is too high to allow the filter to "build up." Ignition pulses, static crashes and flat-topped linears are treated in the same way.

Tune in the Canadian Standard Time Signal or an overseas broadcast (plenty of them on 40 meters). The receiver detector should be on a-m. Notice when the signal fades that sometimes the modulation will become distorted; this is selective fading. Switch in the reciprocating detector and the distortion will disappear.

fig. 2. Reciprocating detector circuit, showing its installation in a Drake R4A communications receiver. Windings of filter FL1 are wound on Ferroxcube pot core and bobbin type 1811CA250-3B7; L1 is 5 turns no. 32 enameled, L2 is 43 turns no. 32 enameled, L3 is 109 turns no. 31 enameled. CT turns L3 to 50 kHz.

Next try CW. Look for the very weak ones and notice that when fades occur on the conventional detector, the signal disappears. Switch to the reciprocating detector, it's still there! This is because the reciprocating detector produces its own beat signal which is proportional to the received signal level. When the conventional detector is on, its bfo level is constant, and so is the low level hiss it produces, acting as a mask for the weak signal.

Sideband operators will appreciate the reciprocating detector because adjacent channel signals which chop up a QSO because they are near the i-f passband are now subdued. Flat-topped linears and lightning noises are almost eliminated by the reciprocating detector; the latter will be completely out of the picture if the noise blanker is used as well.

summary

This unique circuit is well worth the work that has gone into its installation. It is hoped that other amateurs will try it and perhaps find some features we missed; or try to shoot down those reported. To Steve Badessa goes my thanks for the circuitry and his help in incorporating it in my receiver. To my wife, WA1IKR, who typed and typed and typed, many thanks.

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ham radio

monitoring ssb signals

Ssb signal reports based on monitor scopes are misleading if the signal is received via typical narrowband i-f strips

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Although available for some time, little information is known to have been published[1] on the requirements for evaluating received ssb signals using oscillographic techniques.

The purpose of this article is two-fold: (a) to dispel the misconception on the part of many amateurs that accurate appraisal of ssb signals is possible when such signals are received on modern equipment with relatively narrow passbands, and (b) to describe a monitoring method using oscillography that yields an accurate representation of the received signal.

A plug-in module is described for use in an empty filter socket in the Collins 75A4 receiver to increase i-f bandwidth so that a true representation of the signal may be observed and analyzed. No other modifications are necessary to the receiver; that is, the existing socket for the mechanical filter is left untouched and the module is merely plugged in — no holes; no rewiring.

The Collins 75S3 receiver also may be used in conjunction with the signal-analysis methods described here. The plug-in module isn't necessary, however i-f transformers T4 and T5 must be tweaked *slightly* to *obtain the required* response. The monitoring method is adaptable to other receivers as well — the only requirement is that their bandwidth pass the signal components necessary for accurate analysis.

bandwidth considerations

According to accepted engineering practice, a bandwidth at least ten times the modulating frequency is necessary to produce a square wave. Such bandwidth allows one to evaluate components of the signal to its tenth harmonic. Relating this fact to popular ssb receiving equipment, which has a passband of 2.1-2.5 kHz, it is obvious that a bandwidth of 21-25 kHz would be necessary to provide meaningful information on ssb signals. Stated differently, a square wave of only 200 Hz would be faithfully reproduced when received through a 2.1-kHz passband.

monitor system

With the above in mind, I decided to

devise a method to accurately evaluate signals received on my Collins 75A4 as well as on the popular 75S3. In the 75A4, a broadband i-f system was installed, consisting of a simple plug-in module (fig. 1). The plug-in module consists of a surplus Vanguard 22-331-331P double-tuned i-f transformer soldered into a Vector type K1.434 9-pin tube socket.

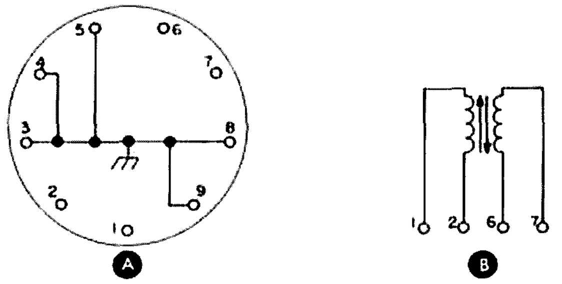


fig. 1. Sketch at A is existing wiring at filter socket XFL in the Collins 75A4 receiver. Broadband i-f transformer, B, may be plugged directly into the filter socket.

The i-f transformer was adjusted for a 20-kHz bandwidth at the 6-dB points using an HP-8601A sweep generator and a Tektronix 7704 scope. With this transformer plugged into a vacant mechanical-filter socket of the 75A4, a broadband monitor system for checking ssb signals is readily available to anyone with an oscilloscope of moderate bandwidth; i. e., 500 kHz or more.

Any small 455-kHz double-tuned i-f transformer may be used, such as the Miller 12-C30. The i-f can should be set up using a sweep generator; however, you can accomplish this on a point-to-point basis with a signal generator and oscilloscope, since the response doesn't have to be perfect.

system constraints

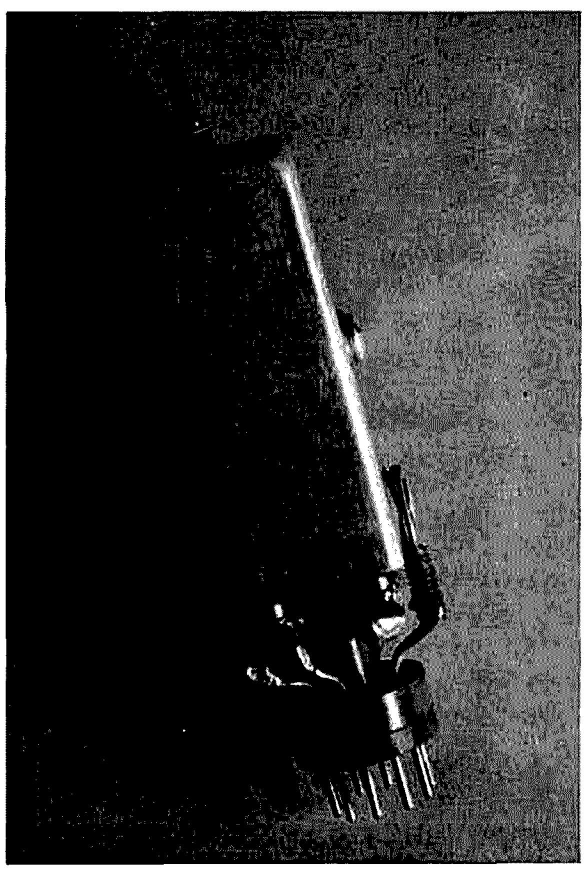
Once the module is installed, the receiver will sound like a party line, with many signals audible, so only the loudest may be evaluated because of the signal-to-noise ratio and interference problems. During my investigation I noted that, in the absence of a mechanical filter, sufficient signal leakage existed around the socket and switch so that very strong signals degraded the true i-f selectivity.

One might immediately ask, "Why do all this when you can look at the output of the second mixer, which feeds the filters, as in most modern receivers?" The reasons are that the signal levels at this point are relatively low; and, more importantly, the reflected input impedance of the filter actually limits the bandwidth so that much of the real value of the system is lost.

In the 75A4, sample the signal to your scope at the plate of V8 with a small capacitor. Retune L27, using the internal 100-kHz internal calibrator. Don't forget to repeak L27 with the system removed, or if the coax isn't terminated.

using the 75S3

Collins 75S3-series owners are fortunate in that the foregoing has been essentially accomplished in the original design. I-f transformers T4, T5 replace a mechanical filter for a-m selectivity. The bandpass of T4, T5 is called out in the



The plug-in module for the 75A4.

instruction book as 5 kHz with unspecified limits. The response of this circuit was measured with a Heath SM105A counter at the receiver vfo, with the S-meter as a readout. The measured response was 5 kHz at the 6-dB points. By readjusting the top and bottom slugs on T4, T5, bandwidth may be extended to 10 kHz at the 6 dB points. This can be done with the aid of a sweep generator or by an approximation using the receiver S-meter and internal crystal calibrator as the signal source.

The 75S3B receiver I checked had almost exactly 6 dB per S-point at S9, as measured using a Kay Lab 6-dB pad. With the 5-kHz bandwidth, in conjunction with a Central Electronics MM-2 monitor scope, absolute correlation was obtained with a slightly flat-topped ssb test signal using an MM-2 scope for direct transmit monitoring.

As expected, no limiting was observed in the 2.1-kHz bandwidth at the same time. From this it is concluded that, while a wideband system is desirable (i. e., more than 5 kHz) it is not mandatory since the observed limiting is relative. This is further substantiated by the fact that, with the relatively narrowband 2.1-kHz filters employed in the transmitters, many of the distortion products will fortunately fall outside the filter passband. Therefore, only the more severe cases of limiting will be observable.

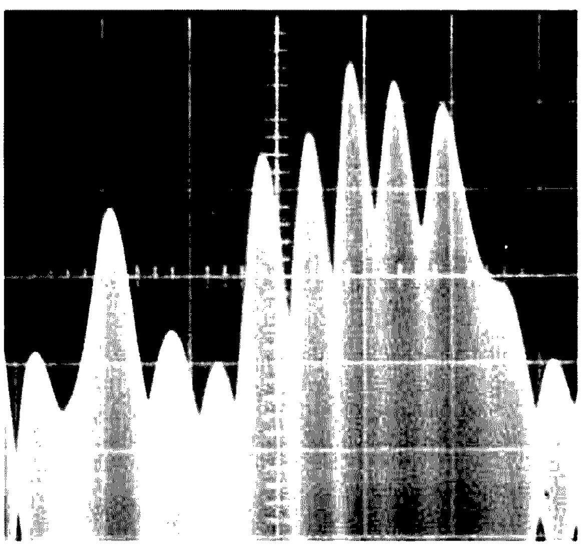


fig. 2. Typical oscillograph display of a clean ssb signal, which was used by author to set up monitoring system.

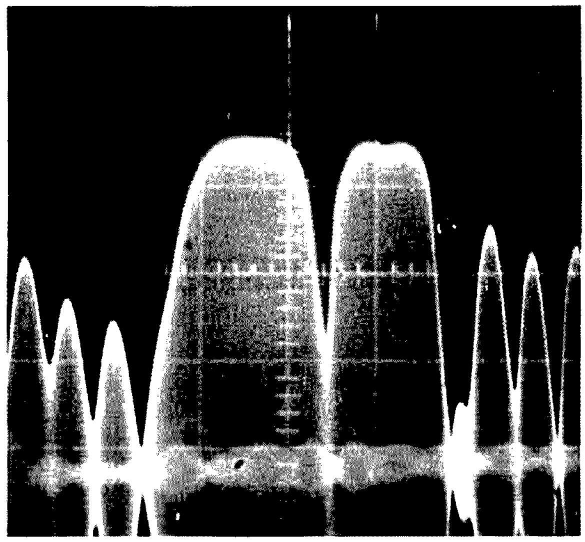


fig. 3. Strong local signal received through 20-kHz-wide i-f strip. Horizontal display is magnified to show detail.

Summarizing, findings with the 75S3B proved that the 5-KHz a-m bandwidth is sufficiently wide to observe moderate limiting. If you wish to see everything, it will be necessary to extend the bandwidth beyond 5 kHz. This is easily accomplished by slightly adjusting the top and bottom slugs on T4, T5 in the 75S3. Sample the signal through a small capacitor at the plate of the i-f amplifier V6 (pin 5) and repeak L9. Turn the *mode switch* to the a-m position. For those wishing to retain the bfo function on ssb, it will be necessary to sample the signal at T4 lug 3. The bandwidth is much greater here, but the signal level is lower.

Other receivers may be used accordingly, bearing in mind that it's necessary to avoid the relatively narrow selectivities, reflected or otherwise, through the filter and its associated i-f system. In any receiver monitor setup, a simple test for adequate bandwidth is that there will be some noise on the pattern, as in fig. 7, and some adjacent signals will be seen that are not audible in the receiver.

test results

Upon completion of the modifications described, you are ready to do some serious looking. The scope display will

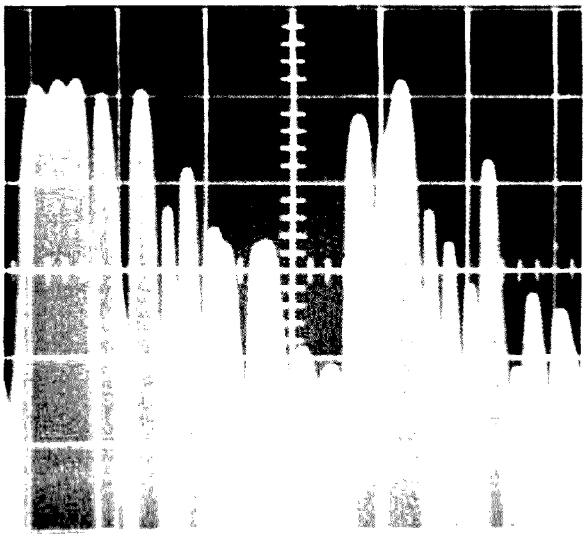


fig. 4. Same signal as in fig. 3. Narrowband (2.1 kHz) filter was used, which presents misleading information as to the true characteristic of a flat-top ssb signal. Compare with fig. 5.

now appear as in a panadaptor. At this point it should be noted that all the photos in this article were candid shots taken with a Polaroid camera mounted on a Tektronix Model 515 oscilloscope.

Tune across the band for a loud signal, turn off the avc and look for a familiar pattern as in fig. 2. This was WA8OWU, who was using an S-Line driving a pair of grounded-grid 4-400s. I just happened to tune across his signal, which is typical of Collins equipment. I used it to set up my scope camera (exposure, etc.). To get this pattern, I used the equivalent of a 30-Hz horizontal sweep rate.

After you become familiar with the system, quickly switch between the 2.1- and the 5-kHz or greater passband, and watch the blinders come off of your eyes! On a linear signal, the peaks will be clean and sharp in either case. Find one that is flat topping on the broadband system and note how it *appears* to look cleaned up in the narrowband system.

A strong local signal was evaluated; see fig. 3. Note that successive peaks begin to decrease in amplitude from the left and moderate-to-heavy limiting exists at the center. Also it should be noted that (a) the scope baseline is near the bottom of the screen and (b) the horizontal display is magnified, providing the fewest possible patterns to permit greatest detail.

In fig. 4 the same signal is shown a few minutes later through the 2.1-kHz filter. This signal appears to be fairly acceptable, because the filter eliminates the harmonics (but not the excessive fundamental bandwidth associated with this condition).

To verify the results of fig. 3 a Tektronix Model 515 scope, which has a 15-MHz bandwidth, was placed directly across the coax transmission line as a broadband receiver. The same local signal was being received. The Model 515 produced a pattern identical to that in fig. 3, which verified that the receiver and the 20-kHz monitor i-f system were doing their job.

In order to verify the linearity of my receiving system, a 14-MHz signal greater than 10^5 μV was injected by a Measurements Corp. Model 80 signal generator into the 75A4 receiver antenna terminals. (The 75A4 had been previously modified with 7360 tubes in the mixers.)[2] Gain was adjusted to provide substantially more than the amplitude on the scope for the setup described, with no limiting. Figs. 4 and 5 are the same signal received on a narrow- and wide-band system, respectively, which demonstrates how a significant loss of information can result.

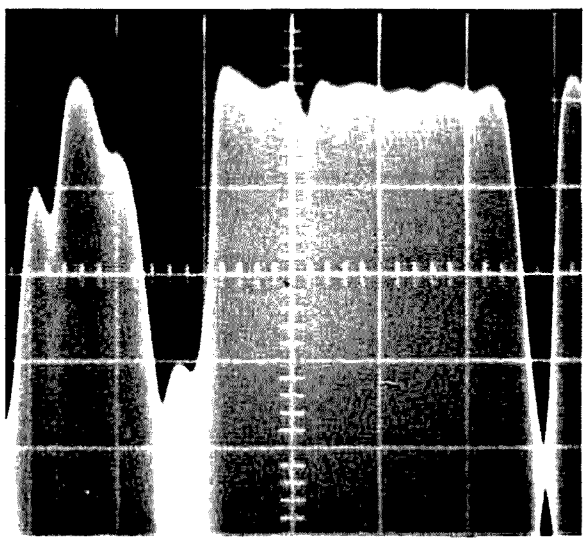


fig. 5. Another wideband presentation of the strong local signal of fig. 3, which demonstrates loss of essential information when received through a narrowband i-f strip, as in fig. 4.

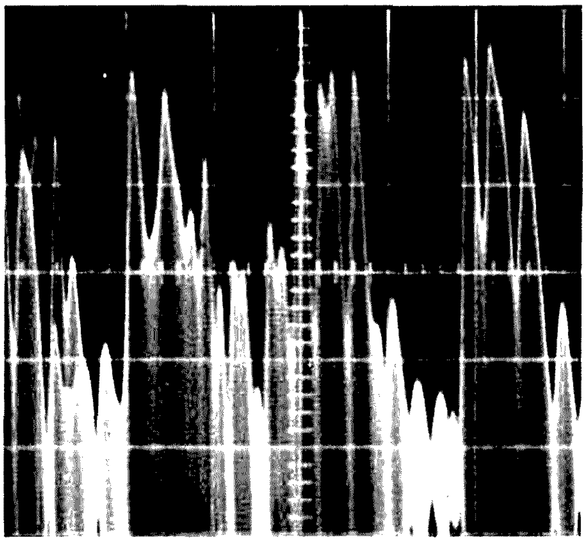


fig. 6. Example of a compressed display resulting from improper horizontal sweep voltage in the oscilloscope. Results are misleading and do not truly represent the signal.

Fig. 6 is an example of a compressed ssb display due to improper horizontal sweep, which also causes misleading information. The objective of the viewer should be to examine only a few highest peaks on the largest and fewest patterns. In other words, you should concentrate only on about one square inch of the scope-tube face.

Fig. 7 is a display of ZS5KI's signal received via long path on 14 MHz, a bit of DX photography I couldn't resist. It demonstrates what a capable system will do under 14-MHz skip conditions. Some noise is in evidence, which accounts for the slightly jagged trace. He used Collins S-Line equipment.

From the above, it may be seen that properly used test equipment is desirable at the receiver and transmitter, especially the latter. No serious ssb station operator should be without a scope to monitor transmitter output. Most any type will do. With direct coupling and a slow sweep, no bandwidth problems exist, and the receivers will take care of themselves.

Most receiver monitor scopes are, in general, improperly used to evaluate ssb signals for linearity when used with narrowband i-f strips. Also most monitor scopes and their receivers probably could be modified to do a proper job if they possess sufficient gain and bandwidth.

In general, linearity is "built into" a transmitter and is governed by the operator. Proper adjustment and choice of operating parameters, such as rf inverse feedback, alc, and rf processing[3,4,5] are among the many well-known techniques for control.

In conclusion, all equipment can be made more linear by reducing power, proper loading and tuning, and by operator-control techniques; i. e., monitoring. These actions must be combined, however. Most importantly, a sincere desire on the part of the operator to exhibit an exemplary signal, rather than the loudest and sometimes broadest signal, is necessary.

I wish to express my appreciation to K6JYO and W6KJD for their assistance in making the 75S3B measurements.

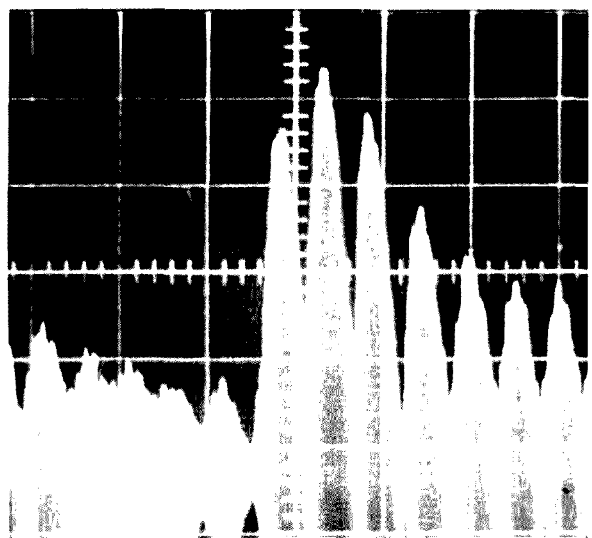


fig. 7. Display of a DX signal received via long path on 14 MHz.

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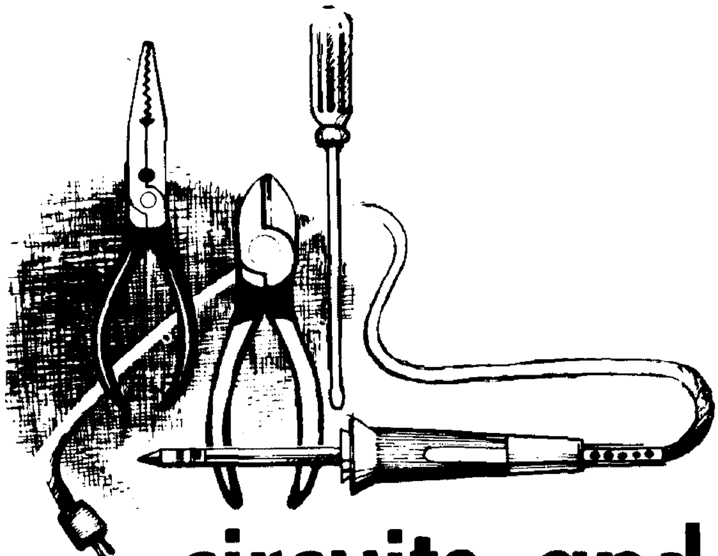


circuits and techniques

ed noll, W3FQJ

digital integrated circuits

Digital ICs are appearing in a host of amateur, commercial and homebrew equipment. Calibrators and counters, indicating instruments, frequency synthesizers, phase-lock loops, ham television circuits, scanners, keyers, RTTY devices, remote control and switching systems, multipliers and dividers, etc. are some applications. Digital devices are entrancing, and delving into then can sharpen your wits and ingenuity. So often we are told with abandon that we need not know any digital concepts nor need any knowledge of what goes on within a complex electronic device. (In fact, modern man is engulfed in a plague of superficial thinking because we are being told continuously we need not understand anything deeply.)

Maybe we can escape for a bit the degenerate philosophy which states, "The only thing important is that which is relevant." Start here by learning a bit about digital concepts even though you don't need to know this to wire a digital device into a circuit and put it in operation. You may even become a bit more conversant with your son's mathematics!

During the next several months I'll gather in some digital fundamentals and show how some simple digital devices carry out these operations. Then you can mount a few of these digital devices on an experiment board and watch their operation with a vom and/or oscilloscope. Lastly, I'll put together some interesting little projects.

counting a new way

Just as there are many languages there is more than one numbering system. Most of us have been hooked thoroughly onto the decimal system. A group of numerals are designated that permit us to count from zero to nine. Then we start over again by placing a one ahead of a zero to give us ten. This will take us up to nineteen; then we start over again by placing a two in front of a zero, etc.

In the binary numbering system, which is adaptable to digital operations, there are only two numerals, 0 and 1. Our brain crevices have become so entrenched that we are astonished to learn we can count with only zero's and one's.

Initially we learn this counting system by associating its concept with the decimal system. We do this in learning languages too. Although we may learn Spanish quite well, our mind does some

fast switchovers between Spanish and our native English tongue. When we really learn Spanish we then begin to think in Spanish. So it is with a new mathematical language. The real digital expert thinks in binary mathematics and in other bases as well. Most of us follow binary with association to the well-worn decimal path.

Nothing in binary and decimal language is zero. Likewise, one apple (now you know when I went to school), is written as 1 in both systems. However, in the binary system two apples suddenly becomes ten. At this point we must make a new crevice in our brain and throw out ten. What we are doing is setting down a one and a zero to indicate in binary form that there is one two (1) and no one (0).

Let's try a three. Are three apples written as eleven? Binary form states that there is (1) two, plus (1) one to represent a three.

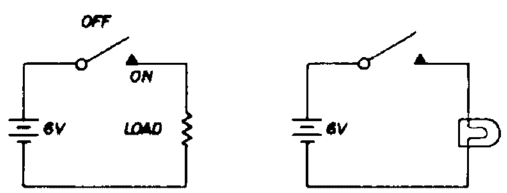


fig. 1 Binary action of a simple switch.

Write a four. You must start out with a new digit. (No wonder we speak of digital concepts and digital circuits and digital computers.) 100 represents four apples is not a true statement. You must think the the (1) represent one four, plus (0) two's, plus (0) one's.

Although a dozen apples is still twelve apples, how would you write it in binary language? The answer is 1100 or (1) eight, plus (1) four, plus (0) two's, plus (0) one's. Table 1 shows the decimal and binary equivalents up to 15. Four binary digits (called bits) are needed to count from zero through fifteen.

The third column is the customary way of writing on the basis of a four-bit presentation. Decimal 1 is equivalent to 0001 in four-bit binary or, (0) eight's, plus (0) four's, plus (0) two's, plus (1) one. The actual count is shown in the fourth column.

table 1. Decimal, binary and code relationships.

decimal	binary	four-bit	weight 8,4,2,1
0	0	0000	(0) 8, (0) 4, (0) 2, (0) 1
1	1	0001	(0) 8, (0) 4, (0) 2, (1) 1
2	10	0010	(0) 8, (0) 4, (1),2, (0) 1
3	11	0011	(0) 8, (0) 4, (1) 2, (1) 1
4	100	0100	(0) 8, (1) 4, (0) 2, (0) 1
5	101	0101	(0) 8, (1) 4, (0) 2, (1) 1
6	110	0110	(0) 8, (1) 4, (1) 2, (0) 1
7	111	0111	(0) 8, (1) 4, (1) 2, (1) 1
8	1000	1000	(1) 8, (0) 4, (0) 2, (0) 1
9	1001	1001	(1) 8, (0) 4, (0) 2, (1) 1
10	1010	1010	(1) 8, (0) 4, (1) 2, (0) 1
11	1011	1011	(1) 8, (0) 4, (1) 2, (1) 1
12	1100	1100	(1) 8, (1) 4, (0) 2, (0) 1
13	1101	1101	(1) 8, (1) 4, (0) 2, (1) 1
14	1110	1110	(1) 8, (1) 4, (1) 2, (0) 1
15	1111	1111	(1) 8, (1) 4, (1) 2, (1) 1

It does not all end here. Higher decimal numbers can be represented using more bits. For example, decimal 30 becomes 11110 or (1) sixteen, plus (1) eight, plus (1) four, plus (1) two, plus (0) one's. Also, more complex four-bit codes can be employed which provide a means of representing higher decimal numbers with special four-bit groups.

binary codes

Digital systems, instruments and devices respond readily to binary information. For example, a simple switch is binary in its activity. The switch of fig. 1, when closed, produces 6 volts across the output. This voltage can be arbitrarily assigned a value of binary (1). When the switch is open there is zero output voltage which can be assigned a binary value of (0). If the load is a small bulb, light *on* becomes binary 1 and light *off* binary 0.

Assume four bulbs are used to obtain a binary representation of a decimal number as in fig. 2. How would you use the switches to indicate decimal number 6 in binary form? The binary value for decimal 6 is 0110. This would be indicated by closing switches 2 and 3. Therefore bulb 1 would be off, bulb 2 on, bulb 3 on and bulb 4 off.

In practical digital equipment the

switches are not manual; they are diodes, transistors or complex groups of switches and other circuits mounted in a digital integrated circuit. In fact, in a practical piece of equipment you would not have to make the conversion between the binary number and its decimal equivalent. This would be done automatically with digital ICs that convert binary information to signals that operate decimal readout devices.

All of these devices respond at high rates of speed that, in a suitable circuit, could read out a very accurate measurement of an incoming radio frequency. In fact, all sorts of quantities and events, regardless of their rate of occurrence can be evaluated with digital instruments.

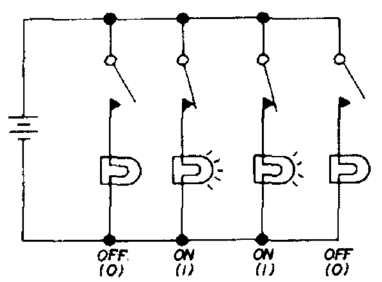


fig. 2. Use of four bulbs and switches to represent a binary number.

High-speed on-off solid-state devices make it all possible.

bcd code

Various codes based on the binary (1) and (0) concept have been evolved to meet the requirements of digital equipment operation and objectives. A pure and simple code which is used extensively is known as the binary-coded-decimal (BCD). Four binary bits are employed. It is said to have a weight of 8, 4, 2, 1 in the order of digits from left to right as shown in table 1 and fig. 2. Each digit position has a definite value (weight). In the pure binary case it is 8, 4, 2 and 1 for a four-bit character. Conversion to decimal values involves simply adding the weights of the digits. For special needs there are various other types of weighted and unweighted codes. Some codes include more than four bits per character.

In the basic BCD code a four-bit number is used to express all decimal signals 0 through 9. Although a four-bit

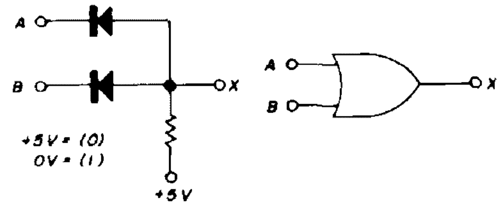


fig. 3. OR-function circuit, negative logic.

number can designate higher numbers (up to 15, as you learned) the BCD code restricts each four-bit character to decimal numbers from 0 through 9.

When a higher number is to be represented in binary form using the BCD code, additional four-bit characters are conveyed. For example, the number 25 in the BCD code becomes 0010, 0101.

2	5
0010	0101

Note the first four-bit character is the binary representation of decimal 2, while the second is the binary representation of decimal 5.

Use the BCD code to write decimal 854.

8	5	4
1000	0101	0100

logic and switching

The closed position of each switch in the circuit of fig. 2 is customarily called the *on* position; open is the *off* position. However, in the language of two-stage or two-logic Boolean algebra (the base upon which computer systems evolved), the closed position could be designated *true* and the open position *false.* True corresponds to binary logic 1; false, to binary logic 0 in switching systems.

In the operation of switching devices there is also an important connection between logic 1 and logic 0 and signal polarity. If the voltage that represents logic 1 (true) is more negative than that which represents logic 0 (false), the

table 2. OR-function truth tables.

chart 1			chart 2		
A	B	X	A	B	X
0	0	0	$\overline{A}$	$\overline{B}$	$\overline{X}$
0	1	1	$\overline{A}$	B	X
1	0	1	A	$\overline{B}$	X
1	1	1	A	B	X

system is said to use negative logic. Conversely, if the signal voltage that represents logic 1 is more positive than that which represents 0, the system uses positive logic. These signal voltages are relative to each other and not necessarily with respect to circuit common (ground).

Often when a negative voltage represents logic 1, it is called a down-level or down-state signal; a positive logic 1, and up-level or up-state signal. Sometimes the terms low and high are used.

OR function

A very basic OR function circuit and its schematic symbol are shown in fig. 3. The two diodes function as switches. When the left side of either switch (diode) is at zero volts, the switch closes (positive voltage on anode and negative voltage on cathode). The output voltage is then zero and the input and output voltages correspond to logic 1. In fact, the output voltage is zero or logic 1 when the voltage at either or both inputs is zero. If the voltage at A and B is +5 volts, both switches are open and the output voltage X equals +5 volts or logic 0.

The characteristics of the OR circuit can be set up in the form of a so-called truth table, as shown in table 2. A truth table is written in terms of logic 1 and logic 0. The first column states that when A and B are at logic 0 potential the output is also logic 0. When logic 1 potential is applied to B, and logic 0 to A,

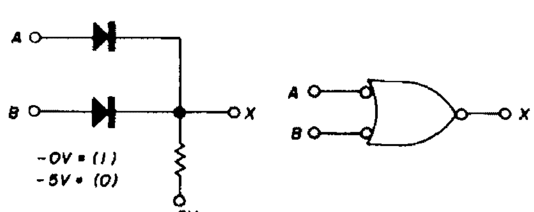


fig. 4. OR-function circuit, positive logic.

the output equals logic 1. Similarly, with logic 1 at A and logic 0 at B, the output is again logic 1. Logic 1 voltage at A and B also results in logic 1 output. Negative logic is used because the logic 1 potential is more negative than the logic 0 potential.

In a logic diagram the actual schematic diagram is not shown. Instead, the corresponding curved line symbol is used to

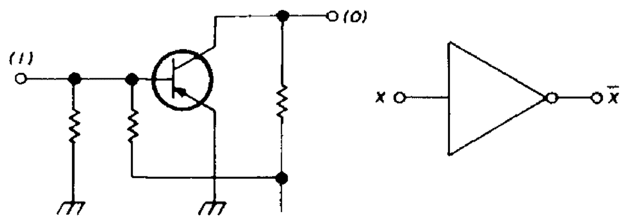


fig. 5. Inverter circuit.

represent a two-input OR-function circuit. This applies regardless of the type of switch – diode, transistor, vacuum tube or integrated circuit.

The OR circuit can also be expressed as a Boolean equation as follows:

$$A + B = X$$

The plus sign in Boolean algebra is not a *plus* but an *OR*. The equation says,

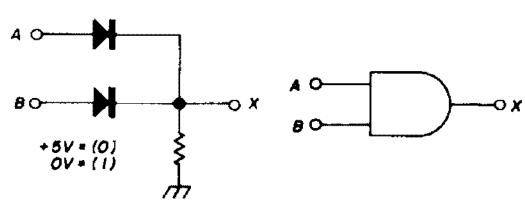


fig. 6. AND-function circuit, negative logic.

"When either A or B (or both) is true, then X is true." The corresponding truth table is shown in table 2B, indicating the same truths as chart 1. A line over the letter indicates false or logic 0. No line indicates true or logic 1. Note that X is true whenever A, B or both are true. The output is false whenever A and B are false.

A similar but positive logic circuit requires a reversal of the diodes and

circuit arrangement, fig. 4. Zero volts is again logic 1. However, logic 0 is −5 volts because positive logic requires that logic 1 be more positive than logic 0. The diode switches close when zero volts is present at A or B. The output is again zero volts or logic 1. A potential of −5 volts at A and B opens both switches and the output is −5 volts or logic 0. The truth table is the same. The symbol is the same except for the bubbles which indicate that logic 1 is more positive than logic 0.

the inverter

Many solid-state circuits, such as a common-emitter stage, result in signal inversion, fig. 5. The inverter symbol is a simple triangle with an appropriate bubble to indicate signal inversion. In this case a logic 1 signal at the input becomes a logic signal 0 at the output; a logic 0 signal at the input, a logic 1 at the output.

AND circuit

In the simple AND-function circuit of fig. 6 negative logic is used with zero logic represented by +5 volts and logic 1 by 0 volts. When logic 1 (0 volts) is present at A and B, neither diode conducts and the output is also logic 1 (0 volts). Both diodes must be nonconducting to obtain a logic 1 output. If logic 0 voltage is applied to A and B both diodes conduct and the output voltage is positive, logic 0. If logic 1 voltage is applied to either A or B the particular diode stops conducting but the opposite one continues to conduct. Therefore the output voltage is again positive or logic 0.

The truth table for the AND-function is shown in table 3. Note that when either or both inputs are of the voltage corre-

table 3. AND-function truth table.

A	B	X
0	0	0
0	1	0
1	0	0
1	1	1

table 4. NOR-function truth table.

A	B	X
0	0	1
0	1	0
1	0	0
1	1	0

sponding to logic 0, the output is logic 0. Only when logic 1 voltage is present at both outputs do you obtain logic 1 output.

The Boolean equation for the AND function is:

$$A \cdot B = X$$

However, the point between A and B is not the ordinary symbol for multiplication but AND. Equation states that, "when A and B are true, then X is true."

NOR and NAND functions

Basically, a NOR-function is an OR-function with inversion. A typical circuit

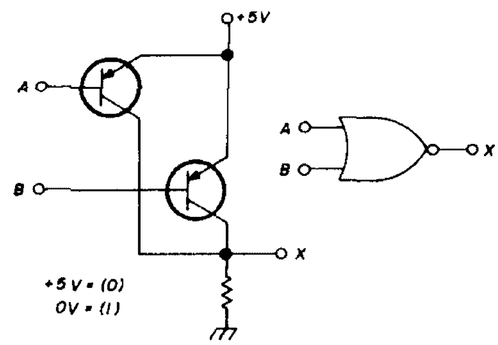


fig. 7. NOR-function circuit.

using the common emitter configuration is shown in fig. 7 along with its symbol; the truth table for the AND circuit is table 4. The common-emitter configuration is a basic inverter because the output is an inverted input signal.

Negative logic is assumed. If zero volts (logic 1 or true) is applied to one or both inputs, one or both of the transistors conduct. At the output a +5 volts (logic 0 or false) signal appears. Note from the truth table for these three outputs there is logic 0.

Refer to the truth table of the OR function (table 2). Observe that for the

same three conditions the output is logic 1 or true. In other words, the output of each is NOT OR which is abbreviated NOR.

When +5 volts (logic 0 or false) is applied to both inputs, both transistors are cut off and the output voltage is zero (logic 1 or true). This too is the inverse of the OR function. Recall that the OR circuit develops a logic 0 output when its two switches are open.

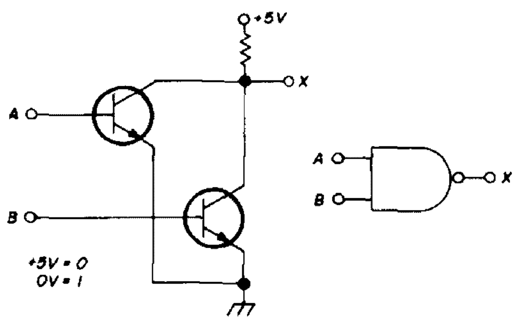


fig. 8. NAND-function circuit.

A NAND circuit is shown in **fig. 8**. In this negative logic arrangement zero volts (logic 1 or true) at both inputs shuts off both transistors and a +5 volts (logic 0 or false) develops at the output. Again there has been an inversion from logic 1 to logic 0. Compare this operation with the AND function circuit of fig. 6. Here again the output is NOT AND which is abbreviated as NAND.

When +5 volts (logic 0 or false) is applied to one or both inputs, the transistors conduct, and the output voltage is zero (logic 1 or true). Reference to the AND truth chart (**table 3**) shows, that with the same input conditions, a logic 0 output is obtained.

Next month I will discuss the special usefulness of the NOR and NAND arrangements and go on to various arrangements of these logic circuits in several basic digital integrated circuits. A bibliography will be given at end of the series.

vhf propagation

The excellent vhf propagation work of R. A. Ham, FRAS, was mentioned in the

table 5. NAND-function truth table.

A	B	X
0	0	1
0	1	1
1	0	1
1	1	0

December column. An article, "The Solar Link," by the same writer in the August issue of "Radio Communication" should be read by every vhf enthusiast and especially by every licensed technician. If you wish to predict openings sooner and add more credibility to your propagation studies, you can find data here.

Mr. Ham uses a solar radio telescope, **fig. 9**, for sporadic-E work. Sounds complex, but it isn't. Solar signals are picked up on 136 MHz using a 4-by-4-element Yagi. A converter moves the incoming signal frequency down to 26 MHz and then into a communications receiver. Detector output is applied to a dc amplifier and pen recorder. Usually solar noise is recorded over the noon hours (1130 to 1330 gmt). Long individual bursts or a continuous noise storm could foretell happy events on 10 and 6, and occasionally 2.

Abnormal tropo reveals its arrival time on Mr. Ham's barometer. Could concerted weather data correlation in U. S. A. give us some more well-defined propagation patterns on 50, 144 and up?

phase-lock thought

Here is an appreciated letter from Thomas H. Morrison, WA3GBU: "In Experiments with Phase-locked Loops" (Circuits and Techniques), *ham radio,* October, 1971, W3FQJ notes that the Signetics 561B IC is not a device applicable to ssb or CW. He is right, of course, regarding this particular chip, but perhaps another chip in this same series deserves a little recognition, especially since it could be useful to those of us who know fm as "chirp." My speculations regarding this IC are not based on personal experimentation, but I feel they are well founded.

The Signetics SE567 chip is a PLL and lock detector in one package. The lock detector is necessary because the PLL itself cannot indicate lock in any way. The typical application of the 567 is as a tone decoder. The brief article, "Need A Tone Decoder?" (*Electronic Design,* October 14, 1971) describes the use of the SE567 in a Touch-Tone decoder, as well as the general principles of tone detection with a PLL.

Only three capacitors and three resistors are needed externally to adjust the center frequency, bandwidth and dc threshold level for the output stage. Center frequency can vary over a wide range, easily covering the normal audio output frequencies of ham receivers. Bandwidth is typically 5 to 10% of the center frequency. Most important is the ability of the PLL to operate in the presence of noise that would make normal LC or active RC networks unusable.

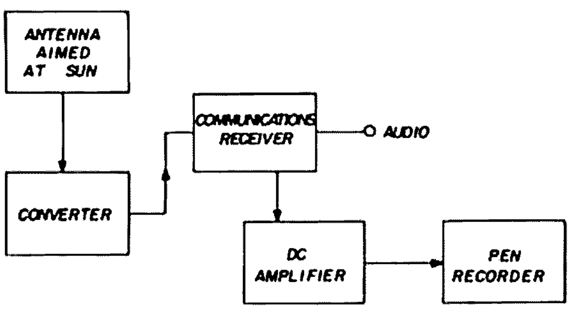


fig. 9. Basic plan of solar radio telescope.

The output of the IC is a logic level indicating lock (tone detected) or no-lock (no tone) which can be used to gate the output of a local audio oscillator (W2EEY, *ham radio,* June, 1970). The user hears the clean, noise-free output of this oscillator (hopefully!).

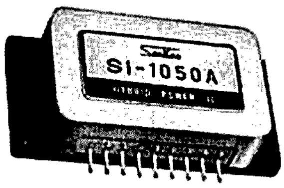
The SE567 costs the same as the SE561. As soon as I can afford the luxury of experimenting with such toys, I will follow up on my speculations. I would be interested in learning of other hams who are using this device.

Thank you, Tom.

ham radio

IMPOSSIBLE? BARGAINS IN SURPLUS ELECTRONICS AND OPTICS

SANKEN HYBRID AUDIO AMPLIFIERS AND SUPPLY KIT



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10 for $ 8.50
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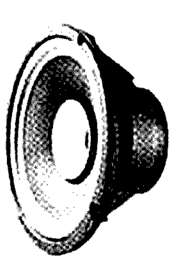
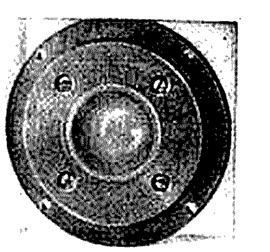
THIS MONTH'S SUPER SPECIAL!

JENSEN HIGH COMPLIANCE SPEAKER SYSTEMS

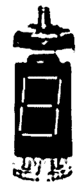




A local manufacturer went out of the speaker enclosure business, and we were lucky enough to buy his inventory of Jensen high-compliance (acoustic suspension) speaker systems. These systems consist of a 12" extended range woofer, a hemispheric dome tweeter, plus crossover. The dome tweeter response extends into the supersonic. The dome shape provides an ideal polar pattern response. The system is ideal for use with our Sanken Amplifier Systems, or any system capable of putting out at least 20 watts rms per channel. Full instructions for cabinet construction are included.

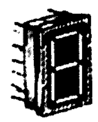
Single System (One Woofer, Tweeter and Crossover) . . $29.00
Shipping weight 10 lbs.
Stereo System (Two of Above) $55.00
Hi Compliance Woofer Only (8 lbs.) $22.00
Dome Tweeter only (3 lbs.) $5.75

7 SEGMENT READOUTS



5 Volt
23 ma/seg.
100K life

A



5 Volt
3 ma/seg.
50K life

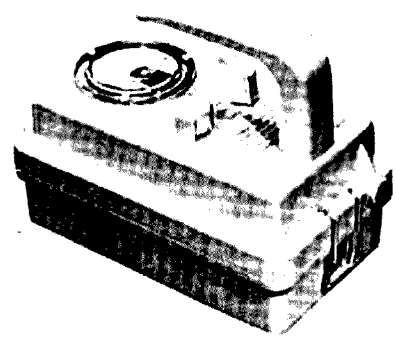
B

7 Segment Readouts. Two types are available, a large size model with wire leads for P.C. Board Mounting illustrated at (A) and a small size low-current version in a Dual In-Line type package for miniature battery operated instruments illustrated at (B).

☐ Large Size Readout (Illus. A) $3.45
☐ Low Current Version (Illus. B) $3.25
☐ Complete counter kit, including 7490 decade counter, 7447 decoder and printed circuit board, and choice of either readout. Price $8.25
☐ Complete counter as above, with 7475 latch, for storage. Price $10.25

☐ Complete Bi-directional counter, with 74192 instead of 7490, for up-down counting $11.25
☐ Complete Bi-directional counter, with latch for storage (74192-7475-7447) $12.25

RADIATION METER ("Geiger Counter")



You can buy a complete radiation meter, complete with original instruction books, at less than the price of the meter movement alone. Range is 0.02 to 50 Roentgens/hour. This is not sensitive enough for prospecting, but useful for other radiation measuring and monitoring purposes. If not used for its original function, then the case, meter and battery holder alone are worth our asking price as a basis for building a metal locator, etc. Uses standard D cell and 22.5 volt Battery.

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More Details? CHECK–OFF Page 94

digital station accessory

Installing the time base, calibrator, identification timer and 24-hour clock

E. H. Conklin, K6KA, Box 1, La Canada, California 91011

Last month I described a number of functions that can be performed by a single digital station accessory sharing a common read-out display. This section will discuss the time-base, digital clock, harmonic calibrator and identification timer. All of these functions are related to a chain of dividers from a highly stable, temperature-controlled, crystal oscillator. In a few cases, no doubt, the power-line frequency will be used until a suitable crystal oscillator is obtained.

power-line frequency

If only for test purposes, in the absence of suitable audio-freqency oscillators, the 60-Hz power line can serve as the time standard. Many circuits have been published for doing this but they appear to be excessively complicated.

When using a 12.6-volt centertapped power transformer and two rectifiers, about seven volts appear from one side of the transformer to ground. Using two resistors in a voltage divider, about two volts rms can be applied directly to the input of a Fairchild 9093 dual JK flip-flop, a SN7490N decade or a Sylvania SM-90 divider (no counting outputs). Some application notes recommend using a diode to remove the negative voltage excursion.

If a bridge rectifier across a 6.3-volt transformer is used, there will be about three volts of ac from one end of the transformer to ground. This will be superimposed upon dc which may prevent toggling the flip-flop. A capacitor in series between the power-supply transformer and the input to the IC will remove the dc. At 60 Hz, the series capacitor can be several microfarads, with its plus lead connected to the trans-

former. A resistance of 1k or more may be placed across the IC so that the input pin will not build up positive voltage sufficient to prevent it from falling to a "logic-low" or "logic-zero" below about 0.8 volts. Frequently, this current-sinking

A JK flip-flop can divide by two, and two FFs can divide by three instead of four without any gates by using feedback. This is done by making use of the Q and not-Q outputs and the J inputs as shown in fig. 1.

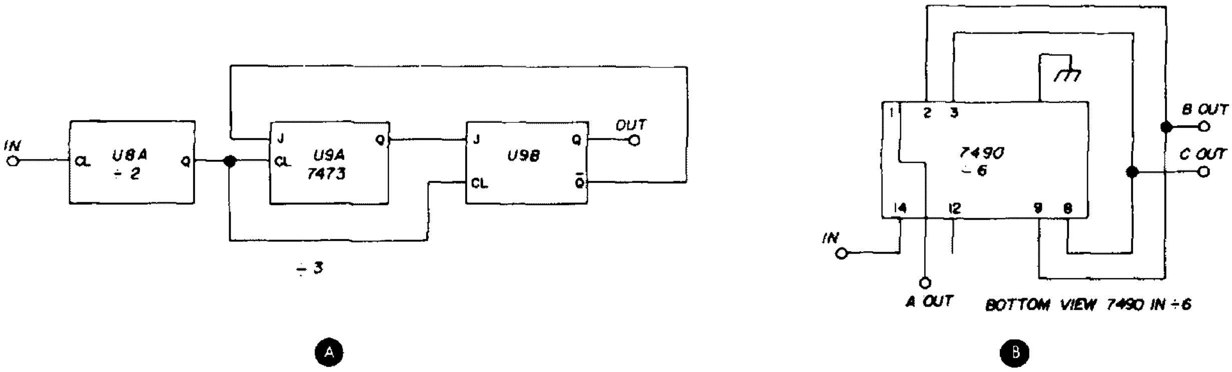

fig. 1. Divide-by-three and divide-by-six using JK flip-flops in feedback method (A), and SN7490N decade counter in reset method (B).

resistor will not be needed. This situation is parallel with "gridleak and condenser" operation of a vacuum-tube detector.

After one toggling flip-flop, almost any FF can follow it. However, the SN7573N does not toggle directly from the 60-Hz sine-wave power source. There are several ways to obtain a divide-by-six, or divide-by-three, using feedback or coincidence gating, so the total division including a decade will be to one cycle per second. **Fig. 1** shows these methods.

The SN7490N can divide by any number from two to ten (except seven) without any external gate. This is done by making use of the two R_0 reset inputs. One can be connected to each 7490 output pin that comprises the binary code for the desired division. The output must be taken at the highest output that toggles. That is, to divide by six, the binary code is 0110. The B and C outputs on pins 8 and 9 should be connected to the two R_0 reset pins, 2 and 3. The C output is taken from pin 8. An external gate will be needed to divide by seven (binary code 0111). This requires the equivalent of three gate inputs to detect the coincidence of a logic-high on the A, B and C outputs.

If the above divide-by-sixty time-base is used from the power frequency, any output can provide a test signal for FFs and for sections of a time-base divider. Should it be used to gate a counter, however, it is usually followed by another divide-by-two flip-flop in order to provide a full one-second *on* time for the count. There are exceptions to this when the *count* and *off* periods are not equal.[1]

crystal oscillator

Crystal oscillators may be ac-coupled (through a stopping capacitor) to the first IC divider. In that case, note that a resistor of 1k to 50k may be required to permit the divider to toggle by providing a drain ("grid leak") from input to ground.

Last month I mentioned Monitor Products Company* ovens and crystal oscillators. Inquiries to Bliley and International did not develop any comparable unit produced for stock. I have a 12-volt version of the SO3233 from Monitor. In normal production this is made for a 30-volt dc supply on the oscillator. It

*Monitor Products Company, 815 Fremont, South Pasadena, California 91030.

would be more convenient to use a 5-volt supply, but then the internal zener would not be able to regulate at five volts so sharing the same voltage regulator with flip-flops may cause slight frequency changes.

Usually these oscillators are dc-coupled and TTL-compatible, which means that they burn out immediately if the output from the oscillator is shorted. Some have been made, upon request, with a stopping capacitor (ac-coupled) output; there may be other ways to prevent accidental burn-out, such as using a series resistor or a "one-input gate" with a resistor from the V_{cc} and a diode, as shown in **fig. 4C**. A good ovenized crystal oscillator should be given full protection.

toggling flip-flops

A number of letters have been received from amateurs who ask for help to make their flip-flops toggle – that is, to flip and flop alternately on a pulsating input. Some comment and suggestions along this line appear to be in order.

It is advisable to test all ICs before installing them; a test socket should be built. Use a high-grade type like the Vector R716-1 16-pin socket. Mount it in P-type plugboard (such as the piece to be sawed off the read-out board). Long wires, machine screws or special Vector pins can be mounted and connected to the socket terminals. Colored short, flexible clip leads can be attached to these. Surplus sockets may lead to erroneous tests results due to contact failure.

In general, finite resistances will be found *primarily* in the V_{cc} to ground pins of digital ICs. When a V_{cc} close to five volts is applied, the unconnected pins mostly will assume a logic-high of about three volts. This is why the J, K and reset pins of a JK flip-flop need not be connected. However, the SN7490N decade counter has reset-9 and reset-0 pins. There are two of each, connected through internal inverting gates. As a result, 7490s should have one R9 pin grounded to permit toggling. If no external gate output is connected to the R_0 pins, one of these, too, must be grounded.

When an external gate output is connected to a 7490 reset, the unused reset pin can be connected to V_{cc}. Then, when the gate is at a logic-low, the IC may toggle; if it is at a logic-high, it will reset. However, if the gate's input is in parallel with another reset, such as from an SN7473N dual JK flip-flop, the voltage from the other IC's reset pin may put the gate at an in-between voltage so that the 7490 will not toggle. This may occur rarely. A resistor of up to 56k, depending on the time constant and frequency, can be connected from the gate input to ground to drain off any voltage from the 7473's reset pin. Operation will become normal.

Interrupted direct current on the input is a poor way to test a flip-flop because of contact bounce. One way to avoid this is to switch the input voltage through a cross-connected pair of gates or inverting amplifiers that hold to one side or the other during contact bounce. Another way is to drive the IC under test from a square wave or, for those FFs that toggle on a sine wave, from an audio or power-line source. Of course, it may also work on rf from a grid-dip oscillator which has sufficient output. (ICs may operate on sine waves at some frequencies and not at others.)

The unused negative excursions of the ac drive should be held down to a small value. It the IC is driven hard, these negative peaks can cause damage. This is prevented by connecting a diode's anode to ground, and cathode to the IC input.

The next step is to ensure a swing in voltage below 0.8 volts (logic-low) in DTL and TTL units, and up above two volts (logic-high). Often, FFs will toggle on much less, but this cannot be guaranteed for all types. A dc component, therefore, can prevent toggling.

If the FF is driven from a very short pulse, a few nanoseconds in length, this pulse can be below specification length and not give the FF time to toggle. In

general this is associated with the maximum toggling frequency of the unit. There are similar limits on the length of preset, clear, transfer and load pulses in some ICs.

U60, may be used after the crystal oscillator; this is to minimize any effect of load upon frequency and to help ensure against shorting the crystal oscillator's output. Once is enough! The gates

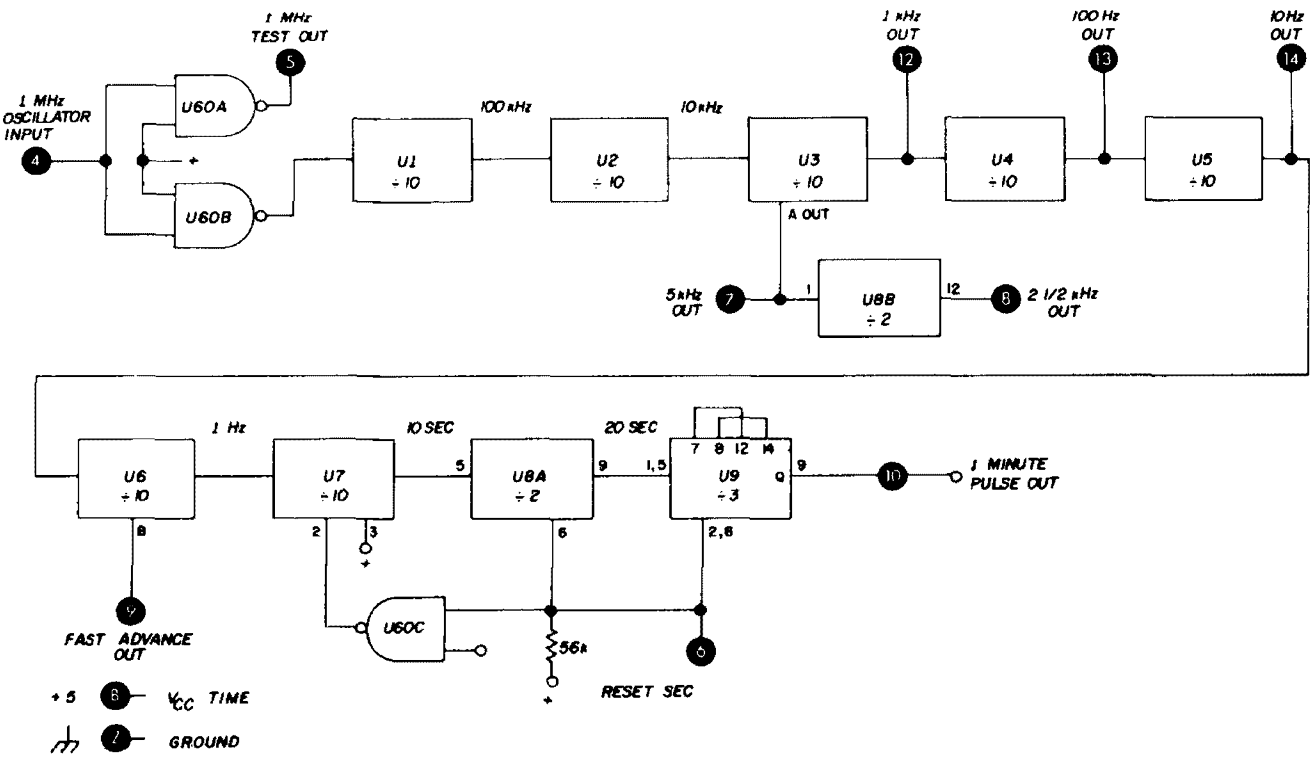

fig. 2. Time-base and clock-driver IC flow chart on the input-time-base board.

Another factor is the *sink* of the input circuit, already mentioned. The plus voltage assumed by some IC inputs must be drained off through the input circuit or the resulting dc build-up will prevent toggling. Overload of the output may have a similar result and may damage the unit if it represents more than a *normal fan-out* given in the specifications. When FFs or gates are ac-coupled to their drive, such as through a stopping capacitor, it is sometimes necessary to use a resistor from input to ground. As already stated, this is a parallel to the old grid-leak-and-condenser tube detector circuit. If the capacitor has a leakage of about 50k ohms there may be no difficulty.

time-base flow chart

Fig. 2 is a flow diagram of the input-time-base board from the crystal input to the one-minute pulse. An isolating gate, feed a string of dividers, mostly SN7490N decades, and also a 1-MHz output phono jack. Three other frequencies are brought out to plugs 12, 13 and 14 for test purposes and to feed a resolution switch for counting, to be described later.

The B output on pin 12 of the sixth decade divider, U6, is fed out to the clock fast-advance pushbuttons PB2 and PB3 to facilitate setting the clock.

An aid to clock setting is a seconds-reset pushbutton, PB1. Although seconds are not displayed at present, they may be, inasmuch as the read-out displays and their decoder-drivers are provided. Even showing only minutes and hours, the seconds-reset is needed so that minutes start in phase with WWV, and do not move up during a specific minute.

The divide-by-six arrangement is the upper one shown in fig. 1A. This was selected to provide the extra FF, U8B, in

the calibrator. Otherwise, an SN7490N would be more simple. With the SN7473s, (U8 and U9), the seconds-reset push-button shorts to ground to reset them, and also feeds an inverting NAND gate to reset the preceeding 7490 decade, U7.

A problem arose here. At a time when the extra V_{cc} filter capacitors were not yet mounted, the minutes output was present but would not reset properly. It was found that the 7490, U7, was always in the reset condition, and spurious peaks in the V_{cc} line caused the next FF to toggle without an input. The cause of the reset condition was an indefinite voltage from the reset pins of the 7473 FFs, U8 and U9. When the latter pins were provided with a sink to ground of up to 56k ohms, the inverting gate, U60C, operated normally and allowed U7 to toggle and reset from the seconds-reset pushbutton.

Inasmuch as divide-by-three outputs can be taken from either the Q or not-Q output of U9A and U9B, the seconds-reset may not recur at the same part of, the minute at which the button was pushed. The output must be taken from the unused Q output shown in fig. 1A in order to have the minutes display change later at the correct second.

calibrator

A 5-kHz output is taken from pin 12 (A-out) of the third decade, U3, for crystal adjustment, for receiver calibrator purposes, and also to feed an otherwise unused FF, U8B, to provide 2.5-kHz calibrator points. When using an audio oscillator and a counter these points are desirable in a ssb receiver to avoid complicating the calculation of frequency by going to lower sideband to hear both the signal and the calibrator. These outputs feed a phono jack on the rear apron of the chassis through the timer panel switch, TS1.

count board

The original idea was to keep similar functions on the same board. This ran into a plug shortage that would increase expense. The solution was to have the time-base to the minute output, the input amplifier, the gating circuitry and the first two count decades on the input-time-base board. Then, the remaining clock circuitry, ID timer, AND-gate switching and four count decades with their latches were placed on the count board, as shown in fig. 3. The final two (megahertz) counters and latches were moved to the read-out board. The advantages of this arrangement will become clear later.

The minute pulses from U9 on the IT board go to its plug 10, and then through a stopping capacitor to the spdt minute-adjust pushbutton, PB2, and on to plug N of the C board. Capacitors from .001 up to 15 μF have been tried, with the large values giving slightly better freedom from irregular contact-bounce pulses when releasing the fast-advance pushbuttons PB2 and PB3. Plug N feeds another stopping capacitor and the first minutes decade, U10.

This in turn feeds U11, a 7460 connected to divide by six, as shown in fig. 1B. Note that the hour-pulse output is taken from C, on pin 8, inasmuch as no pulses reach the D output on pin 11. The output goes to plug T and out to the hours-adjust pushbutton, PB3, then back to the C board through plug U. Another stopping capacitor feeds the unit-hours decade, U12, then a dual SN7473N FF, U13, which will count to four. This is sufficient for 24-hour time.

Inasmuch as the hour dividers must reset at 24 and return to 00, the units-hours 7490, U12, has its two R_0 reset inputs connected to its C output, and to the B output of the two FFs, U13B. The reset inputs are also fed to inverting gate U7OA, so that the number 24 will cause the output of gate U7OA to fall, resetting the 7473, U13, as well. This arrangement did not require any sinking resistor as did the seconds-gate, U60C, on the IT board.

time output circuits

It is desirable to retain the time accurately when the read-out is in use for counting. To do this the V_{cc} must remain

on U10 through U13 and their associated gates. Therefore, the 13 needed outputs are brought to the inputs of four SN7408N quad 2-input AND gates, U71-U74, for isolation. These list at $1.08 each. (The announced SN74157N quadruple 2-line-to-1-line data selectors were not yet available at this writing.) The low-power SN74L98N 4-bit data selector/storage registers are very expensive.

Resistance measurements showed a high resistance from the SN7475N latch outputs to ground in the absence of V_{cc}, so it was decided that the switching between time and count would be by a "wired or" between the AND gate outputs and the latch outputs from the counting decades. "Wired or" means to solder them together and allow a logic-high from either to pass to the read-out board. Later, when using the count mode, the V_{cc} will be switched off the AND gates and on to the latches. At this point in construction, however, neither the AND gates nor the latches are needed. The method requires that only 16 BCD outputs be passed from the C board to the RO board.

id timer

One ID timer on the market carefully provides a coarse and a fine adjustment of the identify time delay for 10-minute accuracy. When the ten-minute signal is produced, it seems to say, "That means that you have just broken the regulations, Bub!" In short, the identification should be made *before* the ten minutes. This may not be convenient if someone else is transmitting. The answer seems to be to use the timer for 9, 8 or some other number of minutes or fractions below a full ten minutes.

The timer should be resetable, to start a new period when the identification has been made. This reset must not change the clock setting. This requires at least one separate IC, U14, driven by a one-minute (or shorter) interval in the timing chain.

The easy figure is eight minutes, by feeding a .7490 in parallel with U10's input, pin 14. However, the seconds

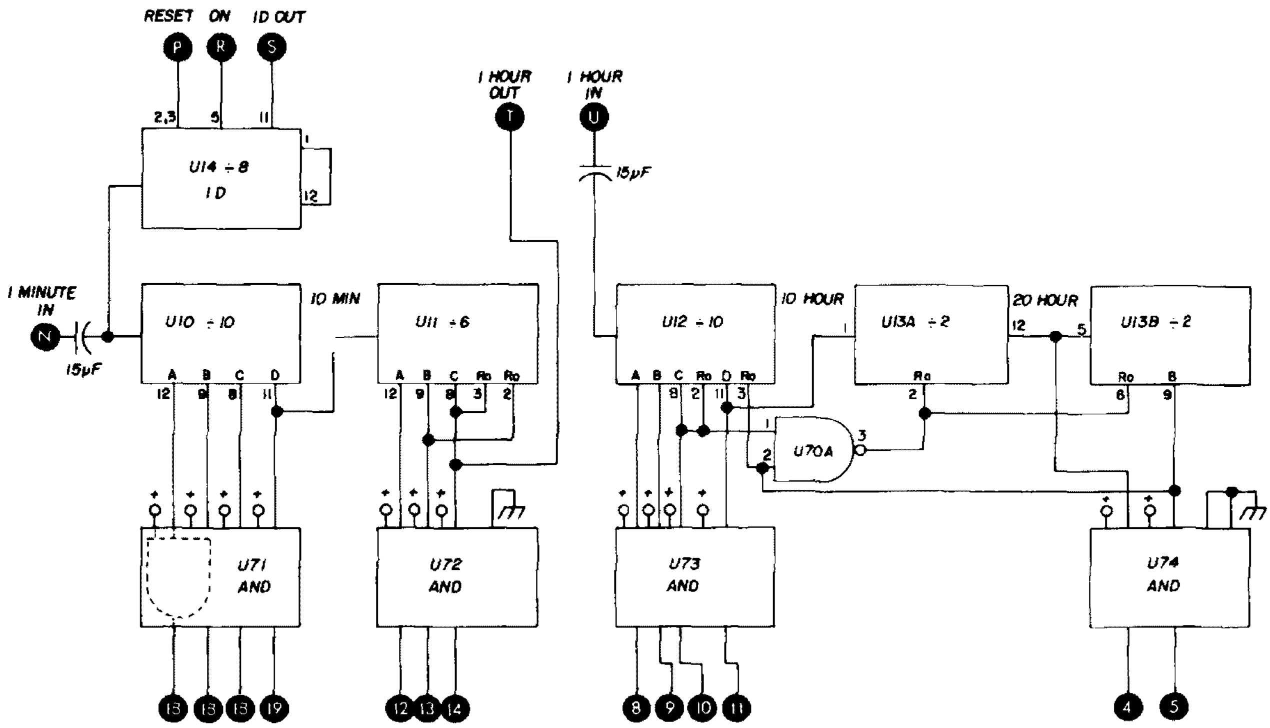

fig. 3. Identification timer, clock ICs and switching AND gates on count board.

status of U9 at the time of reset will differ, making the period eight to nine minutes. If greater accuracy in timer operation is desired, an additional line can be brought from the seconds divider on the IT board, or the timer can be moved entirely to that board. Then, a resetable decade divider can be run from the input to the seconds-dividers. Adding requires a three-input gate (7410). Alternately, several gates can be installed, and the outputs switched. If this switch is set any time during the month to match the number of days in the month, the date will be accurate.

Note that date and month counters require a preset to 1, not a clear to 0. This can be done. The 9093 dual JK

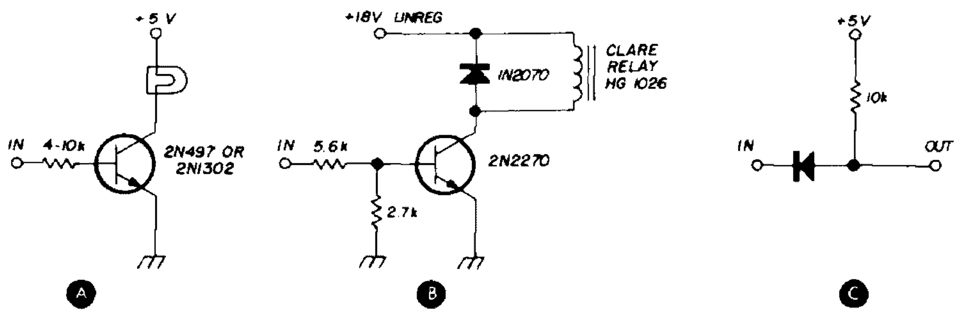

fig. 4. Lamp driver transistor (A). Relay driver transistor with protection from inductive kick (B). Simple one-input AND gate for short-circuit protection of ICs and crystal oscillator (C).

just one decade to the ID timer will reduce the error or variation to a maximum of six seconds, an average of three.

The C output from pin 8 can be fed to a resistor of at least 3.9k, and up to 10k depending on the transistor, which should be an npn device such as 2N497 or 2N1302.[2] The emitter will be grounded. The collector will light a Sylvania 5ESB lamp (49c from Allied or Newark) connected to V_{CC}. See fig. 4. A suitable relay or other device can be driven by the collector current or another transistor. Ac can be used to create a tone in the loudspeaker.

days and months

Ordinarily, it is not desirable to add dates to the time, particularly in areas where power outages are likely to take place. However, a few words on this might bring out some thinking.

By adding a counter driven by the hours output — one pulse per day — the dates will be automatic. This will take a decade counter and a dual JK FF, gated at 32 to preset to 1, for a 31-day month. It is possible to switch the coincidence gate to preset at 31 or 29 days. This presets, rather than clears. Either this or the 7473 dual JK can be reversed by making connections to Q and not-Q (and to J and K if used) in the reverse. That is, a *clear* pulse clears the Q output, but *sets* the not-Q output in the 7473. Both can have one FF preset and the other FF clear. 32 days require a decade and a dual JK FF. One part of the dual JK FF should be used as the first A divider before the BCD part of a 7490, so that it can be preset to 1. The A section of the 7490 can be used with the second half of the dual JK FF, for the tens count to 30.

Counting months is simple with a decade and a JK FF, gated to clear at the count of 13 and preset to 1, which again requires the JK FF to preceed the BCD divide-by-five section of a 7490 for the units decade. Then the A section can follow for the tens count.

At this point it becomes easy to set up more gates to provide reminders of various dates during the year, should there be any desire to do so. The gates could drive a suitable alarm, from the starting circuits in fig. 4. It is interesting that the Sylvania 7420 dual 4-input gate has five inputs on one side, which may be very useful in the

more complicated coincidence-gating of reminder dates.

read-out display

Last month I suggested mounting the Minitron displays and their MSD047 Monsanto decoder-drivers (see TI 7447) on a cut-down Vector 3662 *Plugbord*.

To equalize plug requirements between all three boards the last two counters and latches have been placed on the read-out board, along with a 7400 quad 2-input gate to transfer the latches and provide the 7490 reset.

The positioning of the Minitrons is selected to be visible through the panel window. With some adjustment, the *Plugbord* receptacle could have been raised above the chassis with spacers to give complete freedom of positioning of under-chassis controls.

As designed, the chassis requires that the Minitrons (eight of them) should go into the third column of holes from each side of the *Plugbord.* Including contact holes, the bottom and top pins of the Minitrons go into rows 16 and 23. The bottom and top pins of their drivers go into rows 7 and 14. There is one unused column of holes between drivers.

wiring

Except for the special arrangement of the RO board, all input amplifier parts, gates and divider ICs are mounted on two Vector 3682-2 Plugbords. The DIP ICs are all turned in the same direction to avoid wiring errors.

To prevent damage due to accidental reversal of the boards in their receptacles some thought has been given to reserving the reverse plugs from the V_{cc} inputs. This could receive more study. However, mechanical protection is possible except when using the Vector 3690 card extender (and could be added to that) for test purposes.

Inasmuch as the boards must come very close to the chassis near the receptacle, it is necessary only to add an upward extension to prevent their being inserted. A washer, or a V of wire, can be fitted to the top of the board, extending above the board at about the first row of holes. Then, if the board is inverted, it will be raised above the chassis by this washer or wire key and cannot enter the receptacle.

The Vector 3682-2 boards have printed V_{cc} and ground busses suitable for six rows of dual in-line ICs. The clear space near hole rows 1 through 4 can be used for input amplifier parts, and V_{cc} transient filter capacitors. The buses crossing at rows 47-48, are for V_{cc}. These will be straddled by six rows of DIP ICs. The one row at the bottom, hole columns 38-39, can take five ICs. The others can take four conveniently, if needed.

Space is at a premium, particularly near the IC pins. An Ungar 37- or 44-watt iron does the job, especially with the 1/8-inch round iron-clad tip. This tip should be unscrewed occasionally because the working end lasts longer than the threads. Old copper tips cannot compete.

Solder is a problem because of a need for flux with a minimum of solder. This is met with Kester 44 resin-core solder, core 66, 0.025-inch in diameter. It looks like no. 22 wire.

Miniature precision tools from Sears and others are a necessity, particularly semi-flush diagonals and extra-long-nose pliers.

On a few occasions, a Soldapullt desoldering tool made excess solder disappear. This is available from MacDonald & Company, Glendale, California 91204.

For wire, untwisted light Belden 8430-25 phono pick-up arm cable usually is preferred. About the biggest might be no. 26 (Belden 8505), without having it bend the IC pins. An unusual one, however, is the solder-through insulation *Belsol* wire, Vector 2323A-32-2, which permits crossovers without shorts to an unlimited extent without having to strip the wire. It is useful in series connections like the lamp-test, reset and other lines going from one IC to another. It does take a firm application of heat to get through the insulation, which might best be done when not in contact with the IC pin. None, however, have been damaged so far.

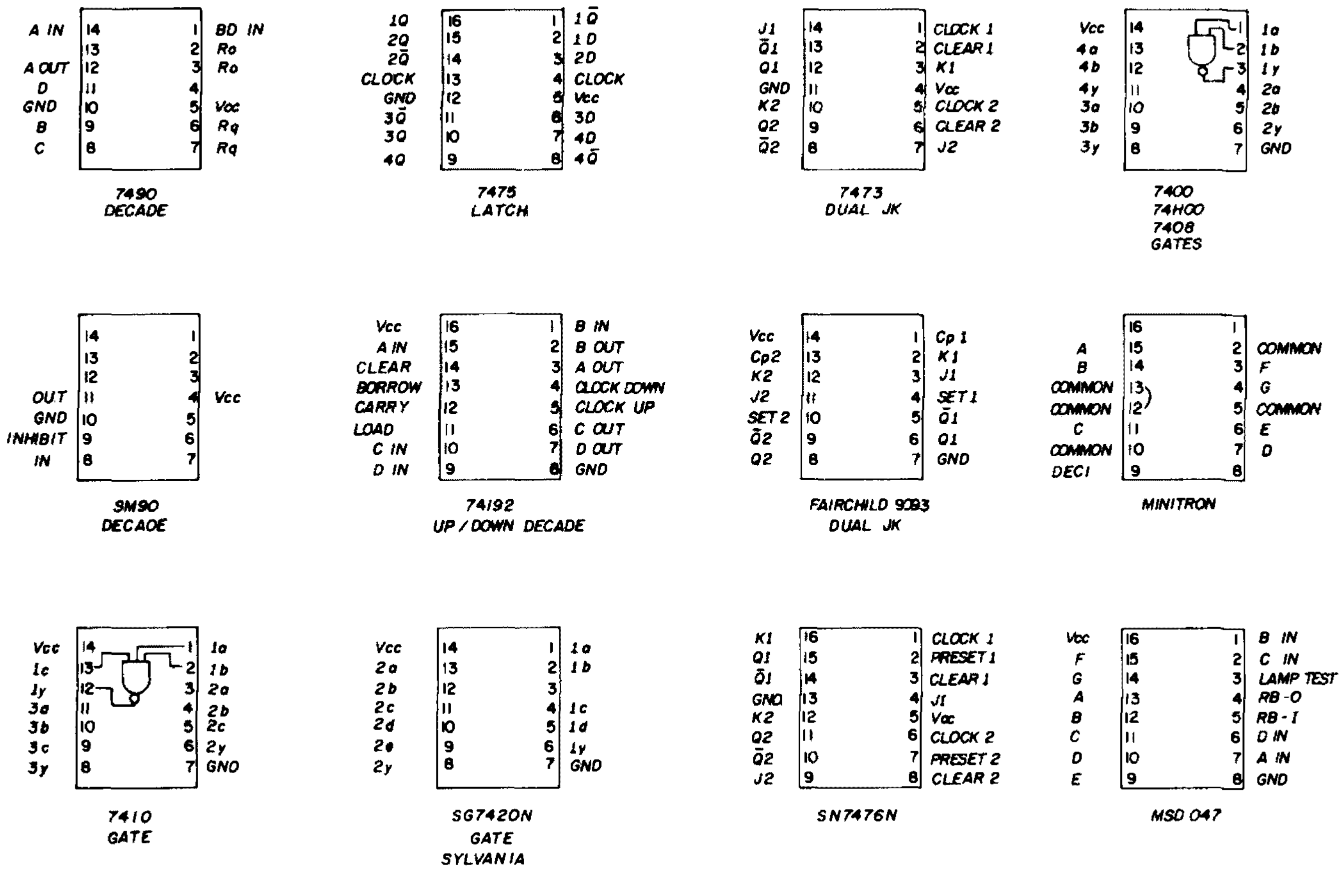

fig. 5. IC base diagrams (bottom view for wiring).

After skinning the plastic-covered wire with a Miller wire-stripper, twisting and tinning, bending a small hook into one end, and putting it on a tinned IC pin makes a good, quick joint. Even three wires can be on one pin.

If several wires are to be soldered to one pin, do them separately. Otherwise the bottom one may not be connected. Test them all with a light pull. After wiring is completed, run a knife point all around each pin to ensure that no stray conducting strand causes a short. Lift the wire from the bus foil and separate adjacent wires to ensure that there is no migration through the insulation due to the heat of soldering.

In the case of the Minitron which may not like to be tinned, it helps to have a loop of wire around the pin and then to solder it. With this procedure no poor contacts have shown up.

Be careful not to put solder bridges across the small space between the pads connecting plugbord contact plugs.

input-time board wiring

On the input-time-base board the crystal oscillator input gate (if used) and the first four decades of the time-base, U1 through U4, can be on the bottom over hole columns 38-39. The second row carries four more time-base ICs, U5 through U8, over hole columns 32-33. Because of using two 7473 dual JK FFs to provide one FF for the 2.5-kHz calibrator output, as well as the divide-by-six seconds-count which could have been accomplished by a single 7490 decade,

table 1. Common plugbord plug assignments. IT is input-time-base board; C is count board; RO is read-out board; TS is the timer switch.

plug	to	purpose
A	IT-C-RO	+5 V_{cc} switched
B	IT-C	+5 V_{cc} continuous
Z	IT-C-RO	-12 ground
Z	TS4B	-5 V
4-7	RO-C	ABCD for U35
8-11	RO-C	ABCD for U34
12-15	RO-C	ABCD for U33
16-19	RO-C	ABCD for U32

table 2. Input-time-base board interconnections. The letter U refers to IC numbers of sections. Small letter P refers to pin numbers.

from	to	purpose
plug 4	U60Ap13	crystal in
U60Ap13	U60Bp9	gate in
U60Ap12	Vcc	unused
U60Ap11	plug 5	1 MHz out
U60Bp10	Vcc	unused
U60Bp8	U1P14	A in
U1p11	U2p14	100 kHz
U2p11	U3p14	10 kHz
U3p12	plug 7	5 kHz
U3p12	U8Bp1	2.5 kHz
U8Bp12	plug 8	2.5 out
U3p11	U4p14	1 kHz
U4p14	plug 12	1 out
U4p11	U5p14	100 Hz
U5p14	plug 13	100 out
U5p11	U6p14	10 Hz
U6p14	plug 14	10 out
U6p9	plug 9	PB2/3
U6p11	U7p14	1 Hz
U7p2	U60Cp3	invert
U7p3	Vcc	reset
U60Cp2	Vcc	unused
U60Cp1	56k R	Vcc
U60Cp1	U8Ap6	reset
U8Ap9	U9Bp6	reset
U9Bp6	U9Ap2	reset
U9Bp2	plug 6	PB1
U7p11	U8Ap5	10 seconds
U8Ap9	U9Bp5	20 seconds
U9Bp5	U9Ap1	x3
U9Ap14	U9Bp8	x3
U9Ap12	U9Bp7	x3
U9Bp9	plug 10	minute

there is one 7473 in the third column over holes 26-27. Lower-numbered columns of holes will be used later for the final time-gating decade for producing count-gate, tranfer, clear or reset and load pulses (for 74192 up/down programmable counters). Note that the negative V_{cc} bus is broken between the top three rows and the bottom three rows of ICs (with Vector line-cutting chisel, P139, $1.50).

General wiring instructions for the IT board are: Cut the V_{cc} bus at row 47, column 24; jumper bottom of V_{cc} bus to plug B. Jumper plug Z to ground-bus. Cut the edge ground bus at row 49, column 1, for 12-V use. Connect all V_{cc} pins to V_{cc} bus. Connect all *ground* pins to ground bus. Solder one input of all unused NAND gates to V_{cc} bus. Solder one pin of all R_9 resets of 7490s to ground. On all 7490s having no gate output to a reset pin, U1 through U6, connect one R_0 pin to the ground-bus. Connect the upper V_{cc} bus to plug A. Connect the long edge ground bus to ground bus (plug Z). On all 7490s, U1 through U7, connect A-out pin 12 to BCD-in pin 1. Further wiring details are given in table 2; plug assignments are in tables 1 and 3.

count board wiring

On the count board U70 provides for 24-hour clock gating and inverting the reset. Its remaining gates are reserved for the transfer signal to the latches. U14 is for ID-timer use. U10-U13 are the minutes and hours counters. U71-U74 are SN7408N AND gates to be used later in switching between clock and count modes. The top two rows are dividers and latches for the count mode. Note that there are two cuts in the connecting V_{cc} buses to divide the ICs into three groups of two rows. This facilitates leaving V_{cc} on the clock counters while the AND gates can be turned on for clock display and off for count display.

Series capacitors, to take dc off the clock fast-adjust lines, are mounted on the count board. If necessary, a few 50k resistors can be added if needed to sink voltages on gate inputs and on stopping capacitors.

Because TTL ICs have noticeable spikes that can trigger unintended ICs, V_{cc} filtering to ground can be placed about every four or five ICs. These can range from discs to substantial 6-V electrolytics from 10 to 100 μF.

The count board requires the follow-

table 3. Input-time-base board plug assignments.

plug	to	purpose
4	crystal	oscillator in
5	phono	1 MHz test
6	PB1	seconds reset
7	TS1D	5 kHz calibrator
8	TS1E	2.5 kHz calibrator
9	PB2-3	fast adj. out
10	PB2	1 minute out

table 4. Count board interconnections.

from	to	purpose
plug N	15 μF C	input
15mfd C	U14p14	timer
U14p5	plug R	Vcc
U14p3	plug P	reset
U14p11	plug S	alarm
U14p14	U10p14	input
U10p12	U71Ap1	A out
U10p9	U71Bp4	B out
U10p8	U71Cp10	C out
U10p11	U71Dp13	D out
U71Ap3	plug 16	A out
U71Bp6	plug 17	B out
U71Cp8	plug 18	C out
U71Dp11	plug 19	D out
U10p11	U11p14	10 minutes
U11p9	U11p3	reset
U11p8	U11p2	reset
U11p2	plug T	1 hour out
Plug U	15 μF C	hour in
15 μF C	U12p14	hour in
U12p8	U12p2	reset
U11p12	U72Ap1	A out
U11p9	U72Bp4	B out
U11p8	U72Cp10	C out
U72Ap3	plug 12	A out
U72Bp6	plug 13	B out
U72Cp8	plug 14	C out
U12p8	U70Ap1	reset
U12p3	U70Ap2	reset
U12p11	U13Ap1	10 hours
U13Ap12	U13Bp5	20 hours
U13Ap2	U13Bp6	reset
U13Bp6	U70Ap3	reset
U13Bp9	U70Ap2	reset
U12p12	U73Ap1	A out
U12p9	U73Bp4	B out
U12p8	U73Cp10	C out
U12p11	U73Dp13	D out
U73Ap3	plug 8	A out
U73Bp6	plug 9	B out
U73Cp8	plug 10	C out
U73Dp11	plug 11	D out
U13Ap12	U74Ap1	A out
U13Bp9	U74Bp4	B out
U74Ap3	plug 4	A out
U74Bp6	plug 5	B out

ing: Cut the V_{cc} bus at row 47, column 18; and again at row 47, column 30. Jumper the bottom of V_{cc} bus to plug B. Jumper plug Z to ground bus. Solder all V_{cc} pins except U14 to V_{cc} bus. Connect all grounds to ground-bus. Solder one input of all unused NAND gates to V_{cc} bus. Solder one input of unused AND gates to ground. Connect the upper V_{cc} bus to plug A. Connect the middle V_{cc} bus, used for AND gates, to plug W. On all 7490s, U14, U10, U11 and U12, connect A-out pin 12 to BCD-in pin 1. Additional instructions are in table 4. Plug assignments are in tables 5 and 6.

read-out board

The middle four Minitrons and drivers, U32-U35 and U40-U43, receive their V_{cc} from the time-base supply, plug B. All others get their V_{cc} from plug A when operating in a counting mode. See pin diagrams in fig. 5 and plug assignments in tables 1 and 6. The Minitron pins 12 or 13 must be connected to V_{cc} and to the other common pins 2, 5 and 10.

All lettered inputs must be connected to the respective lettered outputs of the associated MSD047 drivers. V_{cc} is applied to pin 16 from either plug A or plug B as above. Ground is connected to plug Z. The lamp-test pins are connected together to a wire extending out the back of the board, to which a ground clip can be attached whenever it is desired to display all figure-8s. The ABCD inputs to the decoder/drivers are connected to plugs in accordance with the common assignment table 1.

The blanking of left-edge zeros is a little unclear from some data sheets, but the left-hand driver (ten million), U45, should have its RB-I pin 5 connected to ground, and its RB-O pin 4 connected to the adjacent U44 Minitron's RB-I pin 5. This may be continued to the units digit to the right, if desired. It does not blank the time, because of removal of V_{cc} on the end decoder/drivers. For electronic digital dial purposes, it may be desirable not to blank leading-edge zeros of any digits that normally will be displayed for the method to be used. This will be discussed later.

The detailed wiring instructions for the counter and latch operation for the two left-hand digits will be covered later when I discuss counting.

Keep in mind the relatively heavy V_{cc} and ground conductor currents on this board for the small wire size. Use more than one conductor, entering the ICs at

table 5. Count board separate plug assignments.

plug	to	purpose
B	TS2F	+5 timer
N	PB2	1 minute in
P	PB4	ID reset
R	TS2C	ID V_{cc}
S	alarm	ID timer
T	PB3	1 hour out
U	PB3	1 hour in
W	RS3E	AND V_{cc}

places that will minimize voltage drop.

Sufficient detail is given in the above discussion of the read-out board to wire it without further detailed instructions.

switching

The switch in the upper left position is the *mode* switch (designated MS in interconnection tables), a Centralab PA-1015 4-pole, 11-position non-shorting rotary switch. Tentative future assignments indicate that there is no spare pole which may be needed for up/down count, load pulse control and loading, to be discussed in a future article. Some of the positions may not be used unless the frequency of several five-band receivers is to be indicated, or other types of counting functions are incorporated in the digital accessory.

Below, on the left, is a PA-1005 2-pole, 11-position nonshorting switch. This will be needed to switch to various receiver bands including WWV frequencies (thus it is designated BS) using one pole and over half of the available 11 positions. The second pole, or more, may be necessary for control of up and down counting, load pulse and preset loading in some types of equipment.

On the upper right is the resolution switch (designated RS) which controls displays, presentation of time and decimal-point position. It is a PA-1027 8-pole, 5-position switch; half the poles will be needed and all of the positions. *Off* can be added at the right-hand end of rotation, which can remove the time display while time continues to be available. Some of the spare poles can be used to provide full switching without use of steering diodes.

The final switch, in the lower right-hand position, is a PA-1013 with 4 poles and 5 positions. I plan to use this (designated TS) to control the ID timer, select 5-kHz or 2.5-kHz calibrate signal when desired, and to switch reversed-polarity power to a Palomar electronic keyer currently in use.

The removal of stops should be done accurately, for they probably will fall off if bent back again. Also, note that the switches having fewer than 11 positions, such as the two 5-position switches on the right-hand side of the panel used for resolution, timer and calibrator, are constructed so that it is best to have the *off* position (if any) at the clockwise end of rotation when it is to be separate from five active positions.

Where the usual counterclockwise position is used for *off* the two poles on the same wafer probably will short for an instant. This may not matter unless power is applied by the contact, in which case the short may damage some ICs. This would happen on the timer switch with some types of ICs unless steering diodes are placed in the 5- and 2.5-kHz calibration outputs. Also, it may short the power supply in the polarity-reversing sections if these are on the same wafer.

table 6. Read-out board separate plug assignments.

plug	to	purpose
B	RS1F	+5 display
K	RS2E	decimal U35

A 100-ohm Mallory U1 composition potentiometer, with switch, has been included. This will facilitate adjustment of the calibrator output to match the incoming signal such as WWV, in order to provide a strong beatnote. The 39-ohm resistor shown in **fig. 6** should be increased until the maximum calibrator signal only slightly exceeds the maximum volume of WWV. The switch may be used

later to release some other control, if necessary.

For timer reset, a Switchcraft Littel no. 103 pushbutton, PB4, is mounted on the panel. This type is used also on the chassis for time adjustments. Smaller types are available if desired.

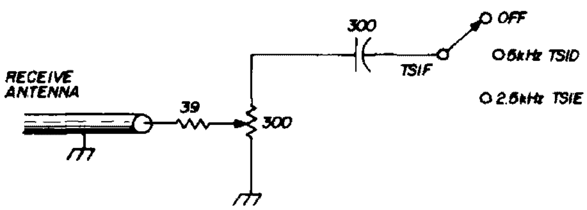

fig. 6. Calibrator output selector. The 39-ohm resistor may be increased to equalize maximum signal with WWV.

Switch contacts in connection **table 7** following numbered designations for each pole, will be given clockwise from the front panel by letters. For example, RS3B indicates the resolution switch, third pole, second contact clockwise from the front. Time-adjust pushbutton circuits are shown in **fig. 7.**

operation

It is well to test the string of dividers on each board on the bench to be sure that they do not present a shorted input or load and that they toggle.

Before the crystal oscillator is plugged in the socket pins should be checked to avoid a short or putting ac on the oscillator input or output; the resistance from the output to all boards should be measured. Then power may be applied to the regulated power supplies. The Minitrons can be tested by grounding the interconnected lamp-test pins with a grounding clip.

Ac and dc tests can be made with a vtvm and scope to show the output at each step in the frequency division. A dc meter usually settles down at about half the V_{cc} (except for the D output from decades) when driven with square waves which are all on one side of the zero line. Be sure that V_{cc} has been switched on the four AND gates for display of time.

When setting the clock, you may reset the seconds (PB1) with WWV even though these are not displayed. They should change the minutes display with the start of the next minute on WWV; if they do not, the output may have been taken from the incorrect divide-by-three output that produces the minute pulse. Next, the minutes fast-advance pushbutton (PB2) should be held down until the minutes are correct, then the button should be lifted promptly. Inasmuch as this count may trigger an increase in the hours display, hours are set last with the hours fast-advance pushbutton (PB3). The hours should pass from 23 to 00 automatically, and be displayed without blanking either of these zeros.

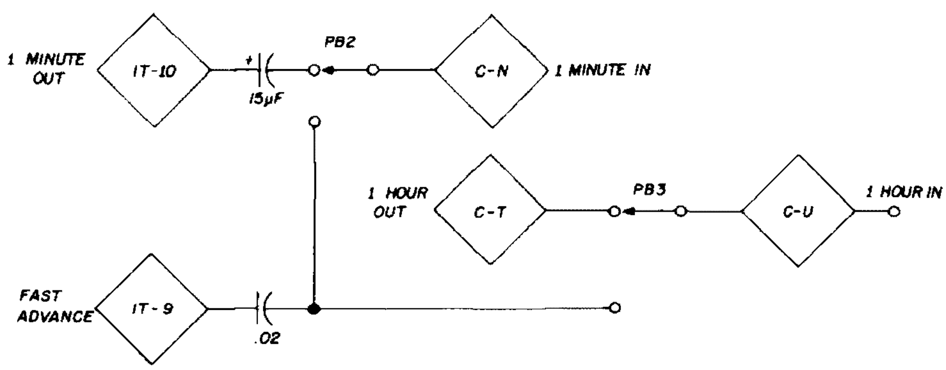

fig. 7. Fast-advance clock pushbutton circuit.

count mode

The next part of this series will cover the input amplifier and count mode using SN7490N decades up to the final digit. Following that, work will be done on the

table 7. Switch interconnections.

lug	to	purpose
RS1B	RO plug B	time readout
RS1B	RS3F	+5 B
RS1C	RS1B	+5 B
RS1D	RS1B	+5 B
RS1E	RS1B	+5 B
RS1F	IT plug B	+5 B
RS2B	RO plug X	decimal
RS2C	RO plug 21	decimal
RS2D	RO plug 22	decimal
RS2E	RO plug K	decimal
RS2F	ground	decimal
RS3E	C plug W	AND +5
RS3F	RS1B	+5 B
RS4B	IT plug 12	1 kHz
RS4C	IT plug 13	100 Hz
RS4D	IT plug 14	10 Hz
RS4F	IT plug 15	resolution
RS5B	TS3B	+5 A
RS5C	RS5B	+5 A
RS5D	RS5B	+5 A
RS5E	RS5B	+5 A
RS5F	plug A	+5 A
TS1D	IT plug 7	5 kHz
TS1E	IT plug 8	2.5 kHz
TS1F	capacitor	receive antenna
TS2C	C plug R	timer
TS2F	C plug B	+5 B
TS3A	phono	+5 keyer
TS3B	RS5B	+5 A
TS3C	TS3B	+5 A
TS3D	TS3B	+5 A
TS3E	TS3B	+5 A
TS3F	+5 V	A supply
TS4A	phono	-5 keyer
TS4B	plug Z	-5 A
TS4C	TS4B	-5 A
TS4D	TS4B	-5 A
TS4E	TS4B	-5 A
TS4F	-5 V	A supply

synthesis method of counting a receiver or exciter frequency with an up-counter, and the use of the 74192 up/down counter with programmed error correction for giving dial indication with a single connection to the receiver or exciter.

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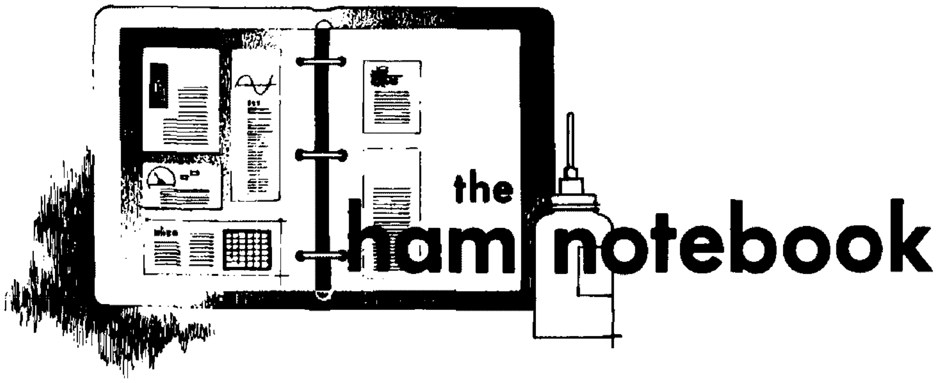
1. Kenneth MacLeish, W7TX, "Frequency Counter for the Amateur Station," *QST*, October, 1970, page 15.
2. E. H. Conklin, K6KA, "Calibrators and Counters," *ham radio*, November, 1968, page 41.

ham radio



testing high-power tubes

Many amateurs would like to take advantage of the vacuum tubes which often appear on the surplus market at very low prices. While a simple emission test, similar to that shown in **fig. 1**, may indicate that the tube has sufficient filament emission and is relatively gas free, something further than such a simple test is needed before attempting to put bargain tubes into service. Most vacuum tube manufacturers test their power tubes with highly sophisticated apparatus which the typical ham can not afford. Very slight leakage can be detected by applying two or three kilovolts on the tube for momentary application and observing what happens. Such a crude test, though effective in some ways, is neither safe for the tube undergoing testing or for the experimenter.

While searching for a better way, the vtvm setup shown in **fig. 2** was developed. The usual grid bias is applied to the tube,

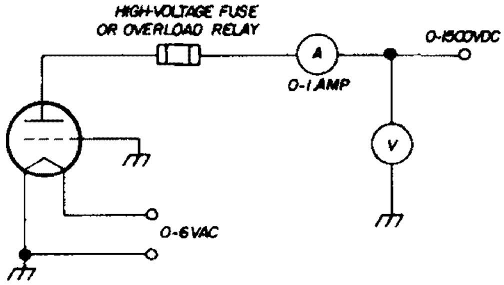


fig. 1. Simple test setup for checking vacuum tube emission.

in addition to the necessary filament supply, and the vtvm – upon appropriate scale – is touched to the anode of the vacuum tube under test. We checked out fourteen tubes with this method, and in every case those tubes which would take normal high voltage in extended service without trouble showed no grid bias on the anode. Those which were excessively gassy showed close to 100% of the bias

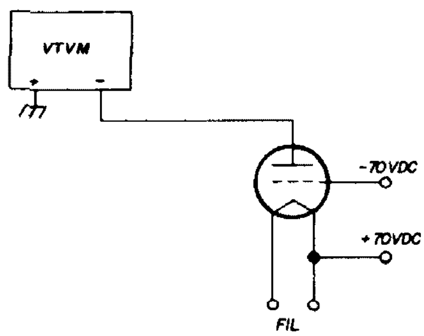


fig. 2. Improved test setup does not use multi-kilovolt power sources. The bias voltage will, of course, vary from tube to tube. The vtvm is set to read negative dc voltage.

voltage whenever the vtvm was connected to the anode of the tube under test.

Obviously, the bias voltage will vary with the type of tube under test, but it is much safer to experiment with 100 volts than to handle, for example, 2,000 volts. This simple method of testing has worked satisfactorily in all cases where triodes had to be checked, and there is no good reason why tetrodes and pentodes could not be similarly tested.

Neil Johnson, W2OLU

diode surge protection

Since building my first power supply using semiconductor rectifiers, I have always been a little uneasy because I never bothered to but in diode surge protection, as in the case of power supplies shown in the *ARRL Handbook.*[1] This bothered me, because I was always uncertain as to the day when I would turn on the power supply and blow the diodes. This has never happened though, and it is a result of this personal experience that I began to investigate surge currents and diodes.

All of the high-voltage transformers I have used have had secondary resistances of between 70 and 300 ohms at voltages from 770- to 1600-volts rms. For example, let us use a transformer model with a peak to peak secondary voltage of 1000 volts and a secondary resistance of 70 ohms. This transformer is fed through a full-wave bridge rectifier to an output capacitance of 60 μF. And for our diode model a set of 1N4007s has been chosen.

At the instant power is applied to the primary of the transformer the load across the secondary represents a dead short (uncharged capacitor). So, until this capacitor is charged, the only current-limiting device in the secondary is the resistance of the secondary winding. For 1000 volts and 70 ohms there is a maximum current flow of 14.28 amperes. And how long this maximum current will flow is determined by the time constant of the secondary (time constant refers to the rate of charge of an RC circuit). For a secondary resistance of 70 ohms and an output capacitance of 60 μF, the time constant is 4.2 milliseconds, and represents the amount of time it will take for the capacitor to charge to 63% of its full value. So what this means then, is that if your diodes can withstand a surge current of approximately 15 amperes for 4.2 ms then they should not be damaged.

Let us examine the 1N4007 diodes with the information shown in *The Semiconductor Data Book* published by Motorola.[2] Here, the maximum surge capability is plotted on a graph of surge current versus number of cycles at 60 Hz, (fig. 3). In this case the graph shows that the diode will withstand safely, at room temperature, a surge current of 50 amperes for one duty cycle, which is 16.6 ms. Obviously then, if the 1N4007 can withstand 50 amperes for a duty cycle of 16.6 ms seconds, it will safely carry a surge current of 14.28 amperes for a period of 4.2 ms. After this initial 4.2 ms the surge current will reduce as the output capacitor approaches full charge voltage.

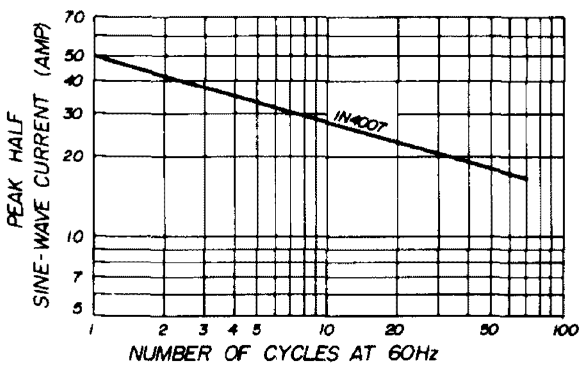


fig. 3. Maximum surge current capability of the 1N4007 diode at room temperature.

Discussions with fellow amateurs about this subject has confirmed my suspicions that surge protection is a needless feature in power supplies. For, of those amateurs spoken to who had built their own supplies, none had bothered to use surge protection. Perhaps this bothersome circuitry was necessary in the days of the top-hat rectifiers; but it now seems to be an anachronism in the days of the epoxy rectifiers.

Here's hoping this article will be of some help to future power-supply builders.

John Lapham, WA7LUJ

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1. "The Radio Amateur's Handbook," 47th edition, ARRL, 1970.
2. "The Semiconductor Data Book," 3rd edition, Motorola, Phoenix, Arizona, 1968.

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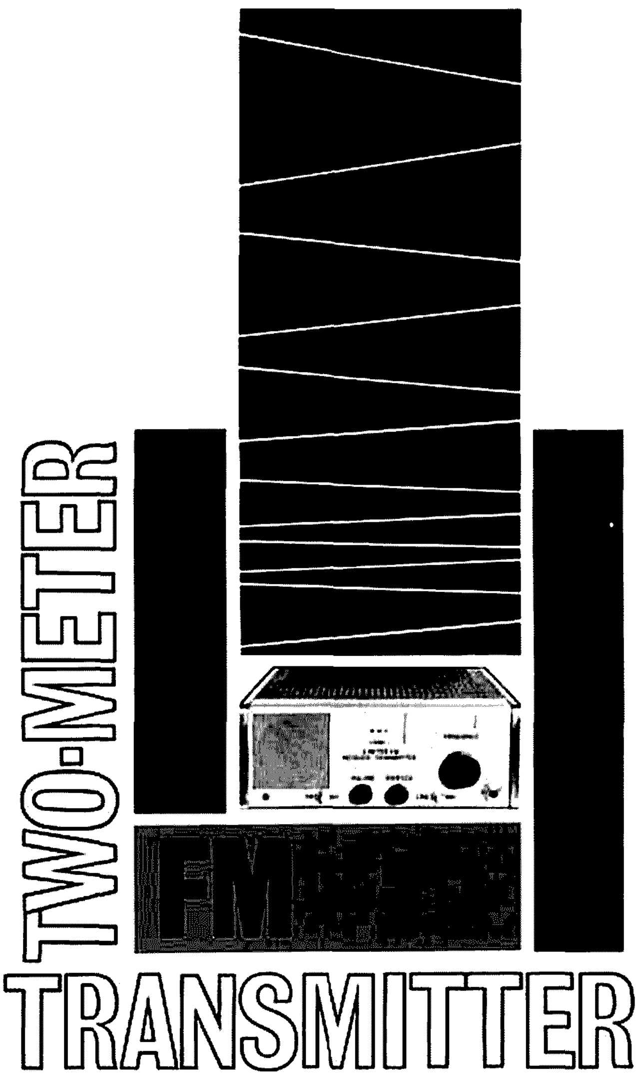
APRIL 1972



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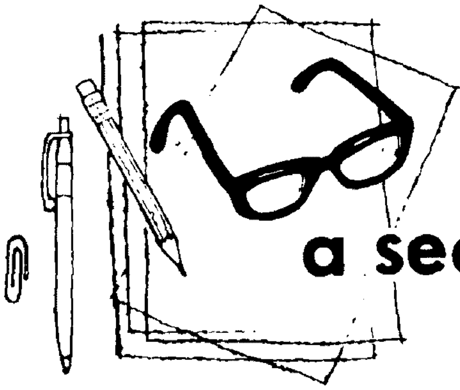


a second look

by jim fisk

Early this month a team of scientists from the Signetics Research and Development Department in Sunnyvale, California announced completion of a revolutionary new *instrument that promises to* transform every field of human endeavor, from radio communications to theology. Led by bio-electronics expert Dr. Ulfias Stopgapski, famous for his theory of positive retrogression, the team unveiled the instrument, called the *omphalometer*, on the first of April.

"The instrument is the ultimate outgrowth," Dr. Stopgapski said, "of my theory of retrogression. I first conceived the idea while assisting my good friend, Professor Ottmar Heissluft, catalog some Tibetan relics in Stanford University's Museum of Anthropology. We were cataloging what were reputed to be dried parts of the antomy of the Yeti, the so-called Abominable Snowman of the Himalaya. These were actually dried navels. I noticed that no two were alike, and this led me to launch a small investigation of my own, involving several volunteers from the Life Sciences Department. Sure enough, all the navels were different."

By using the omphalometer (the name comes from *omphalo,* meaning navel, with the suffix *meter*, or "means for measuring") an operator can perform *omphaloskepsis*. This later technique is, in Dr. Stopgapski's words, "Mediation while gazing at the navel, as practiced by some mystics. This is the first electronic instrument in the world by which the American public can become mystics and omphaloskeptics at home. Monastaries and prolonged meditation are now obsolete."

After sample markets in the United States are tested, the instruments will be exported to the Far East where the demand is expected to be enormous. It may also play a vital role in the Vietnamization Program, although details are reported to be classified.

With the latest scientific evidence showing that palmistry has some validity, the medical world is hailing the omphalometer as a significant new means of prognosis, for it establishes the pattern of the navel's convolutions and thus indicates the future course of a disease. In fact, a patient's *omphalospectograms* can be compared by computer with thousands of others to predict future events and future illnesses. To convert a standard omphalometer to medical use the operator merely attaches an *OmPhallus Probe* to the patient's navel.

Nor is medicine the only application for the versatile omphalometer. One security officer, speaking for the members of his profession, stated that the instrument will soon be installed at all plants that have won contracts from the Government. "It's a natural," he said, "No two belly-buttons are alike, and you know what that can mean to the surveillance and detection people. Why, heck, we'd do away with ID badges. All the employee *would* have to do would be to show his navel to an *OmphaloScanner,* and the door would open electronically, click-shoosh, at least if he didn't have anything to hide."

Jim Fisk, W1DTY
editor

My thanks to Roy Twitty, public relations manager of Signetics Corporation, for providing this bit of nonsense, which combines equal parts of St. Patrick's Day blarney and April First foolery.

two meter fm transmitter

Reliability, economy and proven design characterize this five-watt fm transmitter

Ron Vaceluke, W9SEK, 17 West 540 Hillcrest, Wood Dale, Illinois 60191

In the September, 1970 issue of *ham radio*, Joe Price, WA9CGZ, and I described a 2-meter fm receiver and promised a companion transmitter. This is it! Our goals were similar to those laid down for our receiver: it had to give good performance without costing a fortune to build. Power output had to be adequate not only for repeater use but for point-to-point work as well. This meant that a power output of about 4 to 6 watts was desirable, without the use of vswr protection circuitry.

swr considerations

The lack of vswr protection may raise a few eyebrows, but let's consider this for a while. In the "good old days" when tubes were still being used, vswr protection was never thought of in a ham rig. No one in his right mind would operate his rig without an antenna connected

RMV Electronics, Box 283, Wood Dale, Illinois 60191 can supply a complete parts kit and assembly manual for this transmitter for $59.95 plus $1.40 postage. RMV also can supply just the printed-circuit board, coil forms and assembly manual for $14.50. postage paid. The ferrite beads mentioned in the text are also available, nine for $1.00, postage paid, from the same source. Illinois residents should add the 5% sales tax.

because the tube in the final amplifier would blush red before turning white hot and melting. Then the operator would blush and . . .

Of course, a tube was more forgiving; it could take more abuse before failing than could most transistors. A high vswr is not only rough on the transmitter but is also an indication that something is not performing properly, which usually amounts to poor efficiency. This results in having to run more power output to make up for losses in a poor antenna system. The cheapest way to improve not only transmitter output efficiency, but receiver sensitivity as well, is to use a good antenna with a low vswr. This, of course, holds true for any rig, on any band.

Occasionally things go wrong and damaging conditions can crop up (or *off* when the car wash chops the antenna from the car) which place the final amplifier in jeopardy. Most commercial rigs have vswr protection circuits built in because of economic necessity. After all, it's far better to install an inexpensive rf power transistor and its attendant protection devices than to invest in a good unprotected rf power transistor. The only problem with this is that protection circuits often fail at times when they are needed the most.

The transistor used in our design is made by Fairchild Microwave and Optoelectronics, part number MSA8508. This device was chosen after testing less expensive devices which would blow when the load was removed or the antenna jack was shorted. The MSA8508 withstood this test as well as vswr from unity to infinity.

crystal oscillator

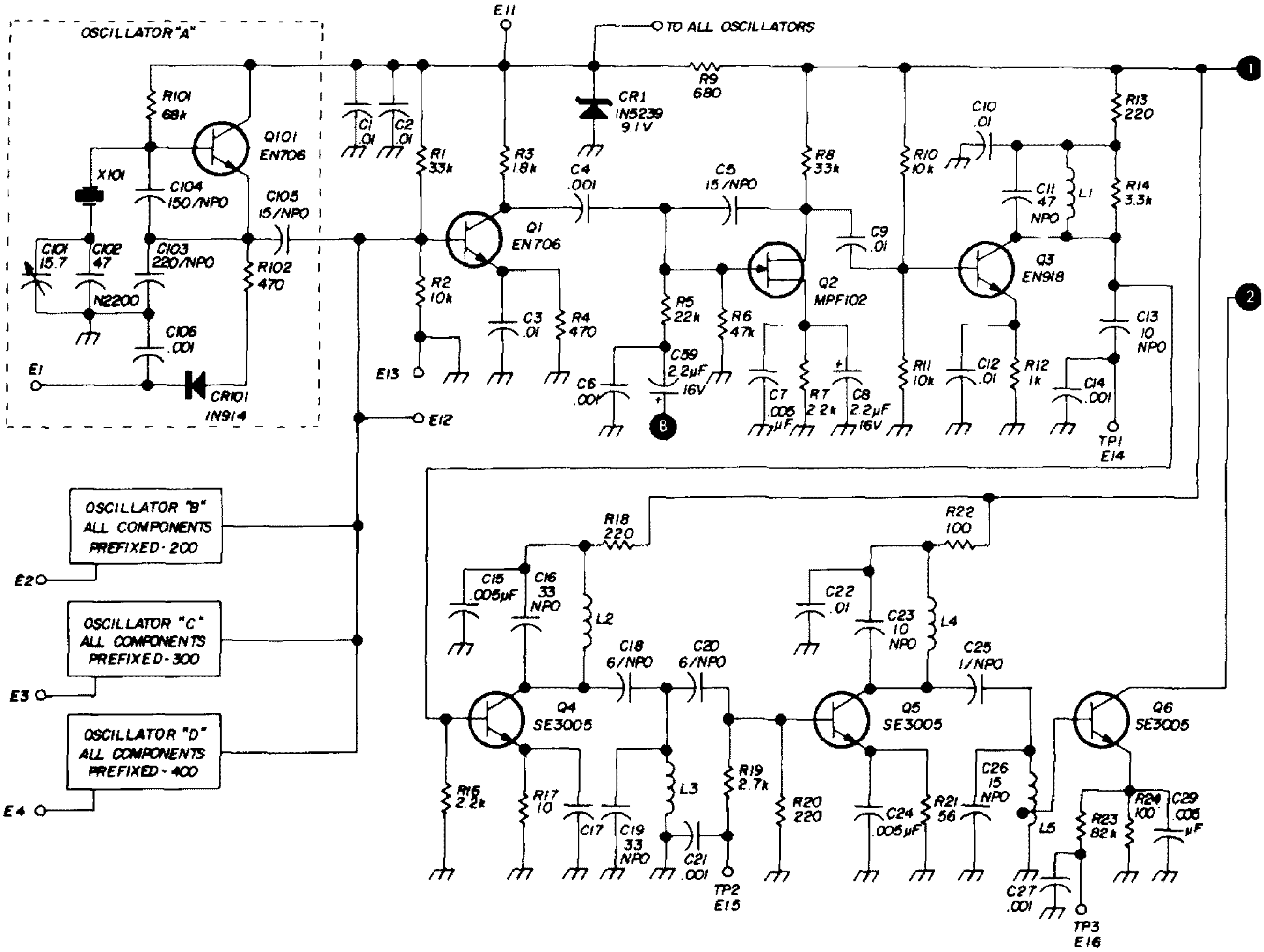
The crystal oscillator was next in line for a shakedown. I felt that an oscillator that could be switched remotely would be more desirable than one which had to have its crystal switch placed right at the transmitter board. Early attempts at switching the crystals with diodes did not work out very well because of the diodes' change of junction capacitance and resistance with change in temperature. Commercial two-way radio equipment has always used a separate oscillator for each frequency, and it's easy to see why. This, then, was our choice also; four channels would be a good start and could be added to, if desired. If maximum stability is desired, use a good quality crystal manufactured by a reputable firm. Buying poor grade crystals is false economy. We recommend the 0.0005% temperature and frequency tolerance; crystals will vary widely from unit to unit even though they are made by the same manufacturer.

circuit description

Each oscillator has its emitter lead brought out to a terminal which is grounded for operation and left open if not used. A diode is included in each emitter lead to provide isolation and prevent false operation if other voltages are present on the frequency-selector switch contacts. The output of each oscillator is connected to the input of a buffer amplifier, Q1, which is slightly forward biased.

Following the buffer is an MPF102 fet phase modulator, Q2. This has both the rf and audio applied to its gate input and produces a phase-modulated signal at the output. A slight amount of a-m is also produced but is removed by the following stages. Transistors Q3 thru Q6 are all frequency doublers which multiply the 9-MHz signal sixteen times to the resultant 144-MHz signal which is applied to Q7 for further amplification. The outputs of Q4, Q5 and Q6 are all double-tuned to reduce the passage of undesired frequencies which are usually present.

Transistor Q7 is a straight-through amplifier and uses a Motorola 2N4427. I emphasize the manufacturer of this device because I found that transistors carrying the same number by a different manufacturer could not produce the output required while the Motorola units did. This conclusion was not reached on



C32, C48, C54, C101, C201, C301, C401	1.9-15.7 pF variable	C43	7-100 pF variable
		C44	37-250 pF variable
C37	8-60 pF variable	L1	21 2/3 turns no. slug tuned. This and all other coils have an inner diameter of 0.215 inches.

fig. 1. Schematic of the two-meter fm transmitter.

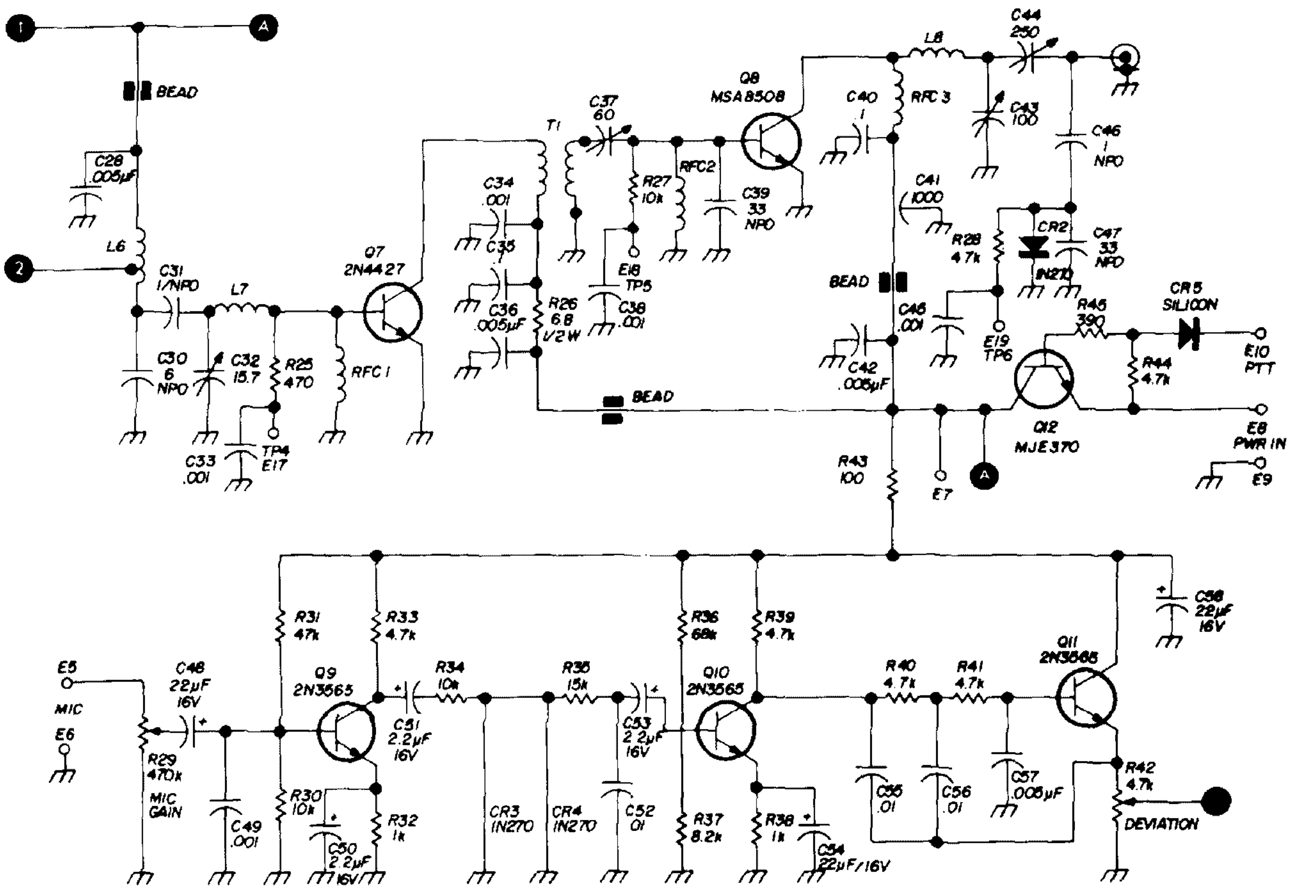
the basis of one rig but of many that have been built. A HEP S3001 is also a good choice for Q7.

The rfcs in the base leads of Q7 and Q8 consist of a 10-ohm resistor and two ferrite beads. The rfc in the base lead of a power-amplifier stage can be the cause of many headaches to someone trying to get one of these devices to work. A choke with a high Q can cause the stage to be highly unstable, while a very low-Q choke will begin to shunt some of the drive and therefore, be wasteful. I found the present value by cut-and-try. The final amplifier, Q8, produces approximately 5- to 6-watts output which can be connected to an antenna or a power amplifier.

Use a 50-µA meter or vom for metering. Test point TP6 uses a capacitive divider and diode rectifier to indicate relative rf output. All other test points *except TP3 measure* the developed base bias of the stage being checked; TP3 measures the emitter voltage of Q6.

audio stages

Transistors Q9 thru Q11 amplify the low output from the microphone to the level necessary for the modulator stage. The input is designed for a dynamic microphone. Two controls are provided in the audio stages. R42 is the deviation control which, when set at maximum, will give about 10-kHz deviation. The microphone gain control, R29, is used to



L2,L3	7 1/3 turns no. 24, slug tuned	L8	2 turns no. 18, air wound
L4	7 1/3 turns no. 22, slug tuned	RFC1,RFC2	2 ferrite beads and a 10 ohm, 1/2 watt resistor
L5	7 1/3 turns no. 22, slug tuned, tapped at 1 2/3 turns	RFC3	5 3/4 turns no. 20, air wound
L6	4 2/3 turns no. 20, air wound	T1	5 1/2 turns no. 20, primary, 1 1/2 turns no. 20, secondary interwound with primary
L7	7 turns no. 20, air wound		

compensate for the output level of the different type microphones that may be used with the rig. Do not use R29 for deviation adjustment because you want as much audio as can be tolerated by Q9. This is because the semilog clipping action of diodes CR3 and CR4 depend upon a strong audio signal being delivered to them. If the output of the microphone, compressor or RTTY afsk generator is very high, Q9 will be driven to the point of distortion; R29 should be set just below this point.

Transistor Q12 switches power on during transmit and off during receive. Normally the transistor is not conducting – it's like a switch that is open. However, when the cathode of diode CR5 is grounded, Q12 will conduct, and the switch is closed. Power is distributed to all circuits through a liberal amount of decoupling circuits. These decoupling circuits are highly recommended to minimize stray coupling through the power distribution circuitry. Part of the decoupling scheme used on Q6, Q7 and Q8 is ferrite beads; they perform this function well at these frequencies.

Power for the oscillators is derived from zener diode CR1 and limiting resistor R9, providing regulated 9.1 volts. Voltage to the transmitter should be kept between 12 and 15 volts. Five-watts output will be easily obtained at 13.6 volts; this is the average voltage found on an automobile battery that is being

charged with a running automobile engine.

The transmitter is built on a 3 x 7-inch double-sided G-10 printed-circuit board. The top surface is used as the ground with all remaining conductor paths on the bottom. Most of the resistors are mounted vertically to save space. If PC-board construction is not desired it can be built on perforated board, but be certain to allow a wide ground path for best results.

Heat sink Q7 with a finned dissipator and bolt Q8 to a piece of 2¾ x 2½ x 1/8-inch aluminum. Heat conduction of these devices to their respective heat sinks is enhanced by the use of Wakefield no. 122 thermal compound which is recommended over the usual silicone grease commonly used a few years ago.

tuneup

Initial tuneup is done in three steps. Power is left off Q7 and Q8 while all remaining stages are tuned. Install a crystal in oscillator A and ground lead E1. Apply +13 volts to lead E8 and -13 volts to any convenient ground point. Measure the voltage at the collector of switching transistor Q12 (point E7) and observe no—or very little (0.1 to 0.2 volts)—voltage. Momentarily ground lead E10 and the voltage out of Q12 should be around 12.5 volts. If the voltage is much lower than this, remove the ground from lead E10 and remove all power from the rig before trying to find out what's wrong. Assuming, however, that the proper voltage is available from Q12, tuneup can begin.

Note that all test points except TP3 produce negative current. Connect a 50-μA meter to TP1 and tune L1 for maximum. Tune L2 and L3 for a maximum reading at TP2 and L4 and L5 for a maximum reading at TP3. Connect the meter to TP4 and tune L6 and C30 for a maximum reading (considerably lower than the readings obtained at the previous test points).

Proceed by connecting power to Q7 and metering the base circuit of Q8 at TP5. Tune C37 for a maximum reading; this should be about 25 μA or more. If everything has functioned well up to this point, the final amplifier, Q8, can now get the smoke test. Connect the output to a load which will present a low vswr at 144 MHz. If a wattmeter is available, connect it to the output circuit. If a wattmeter is not used, connect a meter to TP6. Apply power to all stages and tune C43 and C44 for maximum; power output will be between 6 and 7 watts. Now reduce the output power to 5 watts by using C43 and C44.

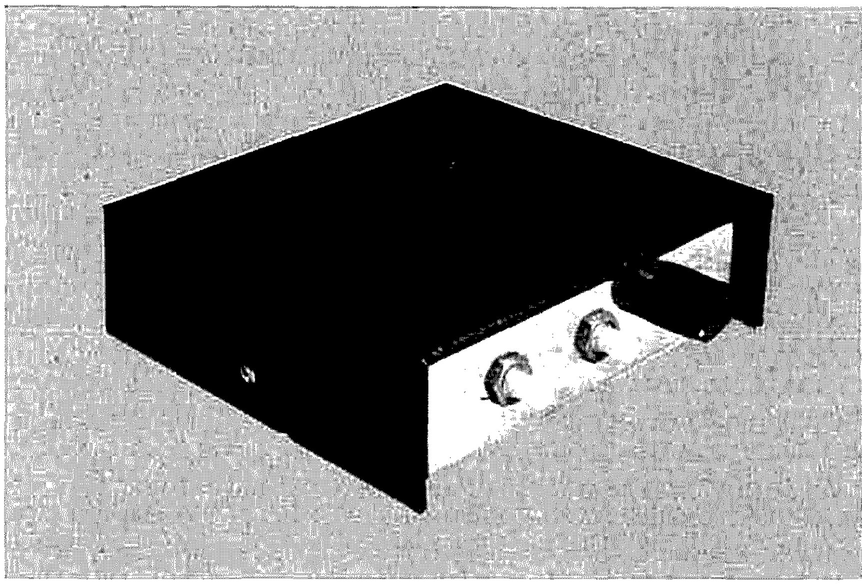
You reduce power because it is better to run the output stage a little cooler than to push for maximum with its higher collector current. The value of collector current will vary under changes of vswr, and some values of vswr can drive the collector current close to the transistor's absolute maximum rating. By backing off on the power output/collector current slightly, the current will never reach this high limit even with a high vswr. It may be wise to meter TP6 continuously when using the rig to ensure proper operation because a high vswr will give a higher reading at this point than is normally obtained under low vswr conditions. Although the final transistor can operate under these adverse conditions, efficiency is poor, and it should be adjusted for maximum performance.

Speaking of performance, always use an antenna which has been cut to frequency by means of an swr bridge and not just by the antenna manufacturer's cutting chart. In the case of mobile installations especially, the location of the antenna can cause the required antenna length to vary quite a bit compared to what the cutting chart shows. If at all possible, use a gain *antenna* rather than a ground plane or a short whip. An antenna with 3-dB gain is the same as doubling your power into a unity-gain device.

reference

1. Ronald Vaceluke, W9SEK, and Joseph Price, WA9CGZ, "An FM Receiver for Two Meters," *ham radio*, September, 1970, page 22.

ham radio



low-distortion two-tone oscillator for ssb testing

A handy piece of test equipment featuring excellent spectral purity

Hank Olson, W6GXN, Box 339, Menlo Park, California 94025

A good sine-wave oscillator is an asset to any ham shack or electronic service bench. For adjustment of single-sideband equipment, one oscillator is a necessity, and two such are very helpful at times. Since most ham shacks are lucky to have even one audio test oscillator, two-tone testing is a rare luxury.

Here is a dual Wien Bridge oscillator that produces 800 Hz and 2000 Hz (both frequencies within the normal voice-frequency range) of good purity for ssb testing. I took advantage of some of the new inexpensive operational amplifier ICs which provide large blocks of gain and ease of application. IC op amps are also used as active low-pass filters to provide optional harmonic reduction of each of the two tones generated. An etched circuit-board layout is provided and laid out so that you can build a simple two-tone generator and combiner with two ICs.*

You can also add one or two stages of

*A printed-circuit board and a complete parts kit for this unit are available from Southwest Technical Products, 219 West Rhapsody, San Antonio, Texas 78216.

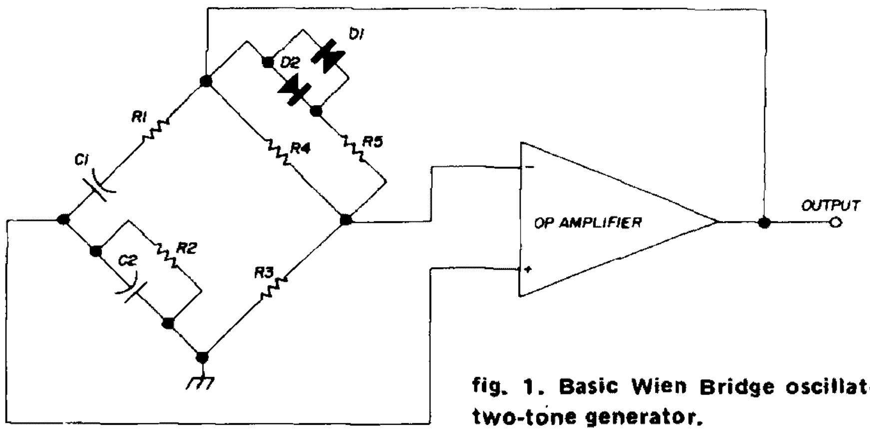


fig. 1. Basic Wien Bridge oscillator used in the two-tone generator.

low-pass active filtering after each oscillator, for successively purer output. The full-blown circuit with two stages of low-pass active filtering following each oscillator provides a two-tone output

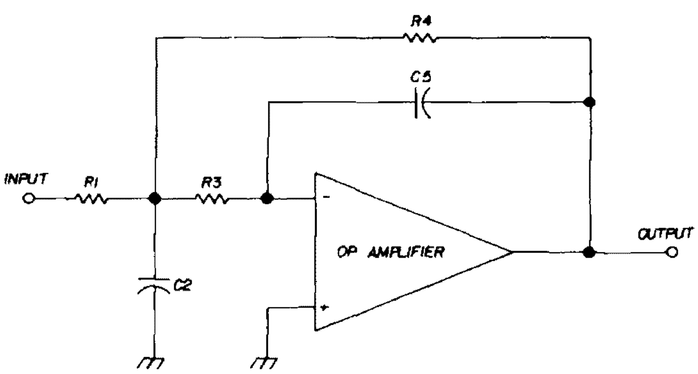


fig. 2. Basic active R-C low-pass filter used in the two-tone generator.

with all other components (harmonics of 800 Hz and 2000 Hz) down more than 70 dB.

The basic Wien Bridge oscillator circuit used in the two-tone oscillator is shown in fig. 1.[1] Note that this variation is different from the classic Wien Bridge oscillator using a lamp as the nonlinear resistance element in the R3 position. In the original lamp version the lamp decreases the gain of the loop by *increasing* resistance with increasing oscillation amplitude. The diodes used as a nonlinear resistance in **fig. 1**, however, *decrease* effective resistance with increasing amplitude; so they are placed in the *upper* leg of the negative-feedback half of the bridge. Like most Wien Bridge oscillators, the frequency of oscillation is $1/2\pi R1C1R2C2$. Unlike most Wien Bridge oscillators, however, there is no feedback time constant (usually provided by a lamp or thermistor *thermal* time constant). The back-to-back diodes operate essentially instantaneously to control amplitude, but since they conduct nearly symmetrically and have a large resistance in series with them, nonlinearity is not too severe. The result of the nonlinearity introduced by the back-to-back diodes is that odd harmonics are larger than ordinarily present in a Wien Bridge oscillator. However, the nonlinearity is not all bad as it helps to increase the stability of the oscillator.[2]

If you are content with a two-tone spectrum that has all harmonics down 40 dB or more, the circuit may consist of only U1 and U4. That is, the circuit is simply the two Wien Bridge oscillators and an operational adder. However, if you want better spectral purity of each of the two tones, follow each oscillator with one or two sections of active low-pass filtering to attenuate harmonics. The

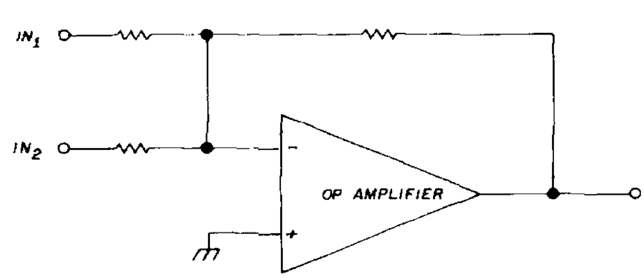


fig. 3. Basic operational adder used in the two-tone oscillator.

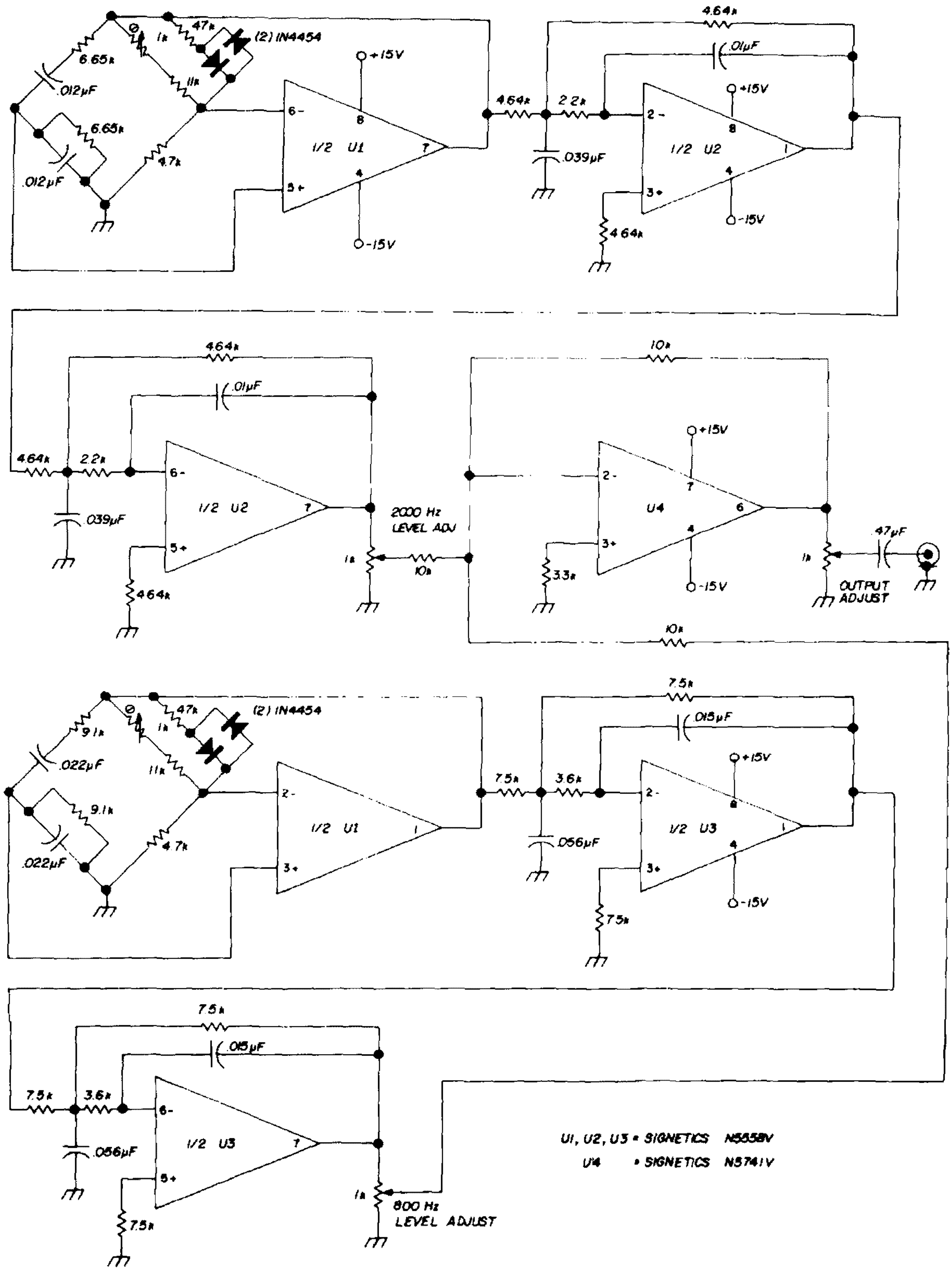


fig. 4. Complete circuit of two Wien Bridge oscillators, two sections of active low-pass filtering for each oscillator and the operational adder.

harmonics of the 800 Hz oscillator are more troublesome, of course, since the second, third, and fourth all fall in what is generally considered to be the audio-frequency range: 300 to 3300 Hz. Depending on your application, the low-pass filter following the 2000-Hz oscillator may or may not be necessary. In any case, because of the way that the circuit board is laid out, the low-pass filters may be left out entirely, one section may follow each oscillator, or two sections

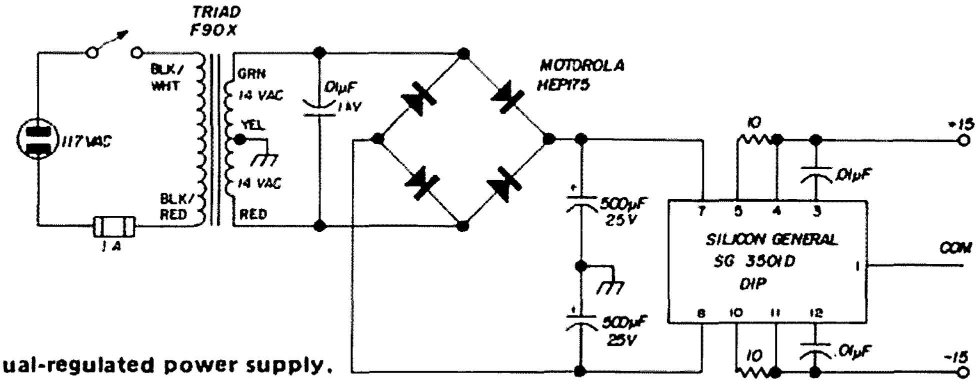
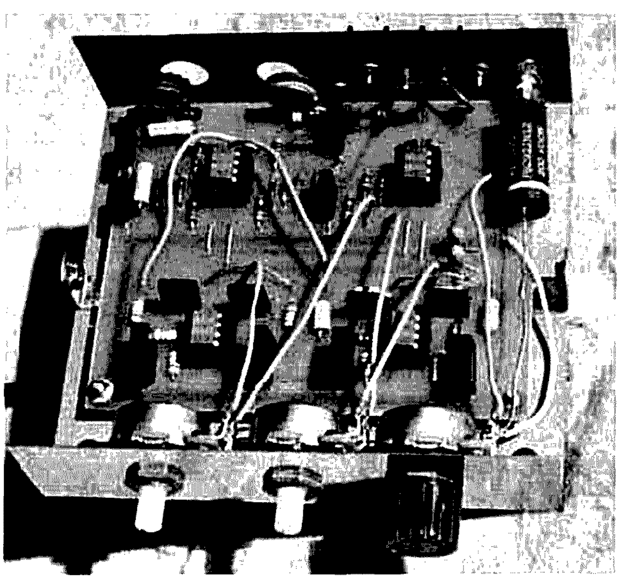


fig. 5. IC dual-regulated power supply.

may follow either or both oscillators.

The low-pass filters, like the Wien Bridge oscillators, utilize IC op amps and R-C elements to accomplish frequency-selective functions. These R-C active low-pass filters are quite stable, easily calculated, and have the advantage of providing gain if desired.[3] The basic active low-pass filter is shown in fig. 2. There are three Rs and two Cs that completely determine the gain and cutoff frequency of this filter. The ratio of R1 and R4 determines the *gain* of the filter (1 in our particular example), but you must not get the impression that varying R1 and/or R4 will affect *only* gain. R1 and R4 also affect the filter frequency cutoff. The cutoff frequency of the filter that follows the 800 Hz oscillator is set at 1000 Hz, and the one that follows the 2000 Hz oscillator is designed for 2500 Hz. Each *section* of active low-pass filter affords a roll-off of 12 dB per octave.



Inside view of the test oscillator shows parts placement.

The operational adder (**fig. 3**) provides a way of combining (or adding) the two tones with virtually no interaction between oscillators. Such an adder is quite often called a mixer in the audio world; but I prefer to call it an adder, because it actually algebraicly *adds* the two inputs. Since the summing point (the inverting input to the op amp) appears to be a virtual ground, each oscillator (or oscillator followed by a filter) sees 10k to ground. This fact, that there is a virtual ground at the very point where the signals are connected together, assures that the two oscillators cannot affect each other. Since the operational adder is a very linear device, it does not cause mixing (why I choose not to call it an audio mixer), and no detectable cross-products are produced. If this adder were nonlinear, of course, we should expect to see all sorts of spurious frequencies such as 2000 - 800 = 1200 Hz, 2000 - (2 x 800) = 400 Hz, and so on. The exact spectrum of spurii would depend on the nature of the nonlinearity.

Fig. 4 shows the complete circuit with two Wien Bridge oscillators, two sections of active low-pass filtering for each oscillator, and the operational adder. At this writing, only Signetics is making available the half dual-inline package N5558V, used as the dual op amp. However, both National Semiconductor and Motorola are soon to offer their own equivalents; the Motorola equivalent is to be called an MC1458CP1. The half dual

inline packaged "741" (Signetics N5741V in **fig. 4**) is also made by Texas Instruments, Motorola National Semiconductor, Fairchild and others. Each of these second-source companies has their own particular (similar) number for the IC. For example, the Motorola number is MC1741CP1.

Note that there are five controls in the circuit. There is a negative feedback control on each oscillator circuit, used to set the amplitude of oscillation. These controls should be adjusted to give about 12 V p-p output from each oscillator (a setting giving too low an output will increase the percentages of harmonics). Since the negative feedback controls are only set once, they are board-mounted screwdriver-adjust trimmer pots. The other three controls — used to set the amplitude of each tone and combined output level — are mounted off the board.

Since the two-tone test generator requires ±15 V at 20 mA, it may be powered by a group of series-wired batteries. Four Burgess F4BP and two Burgess F2BP types are more than adequate. It is a good idea to use a dpst switch with this battery pack so the +15 V and -15 V are applied at one time. An IC dual-regulated supply may also be used. Such a supply is shown in **fig. 6**. The Silicon General SG3501 is used here for both + and - regulation. This IC is only offered by Silicon General at this writing, but is soon to be second-sourced by Motorola.

Measurements of harmonic content (and to check for possible cross-products) were made using a General Radio 1900A wave analyzer, which has a dynamic range of 80 dB.

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1. R. Botos, "Breadboard Techniques for Low Frequency Integrated Circuit Feedback Amplifiers," Motorola Application Note AN271, October, 1966.
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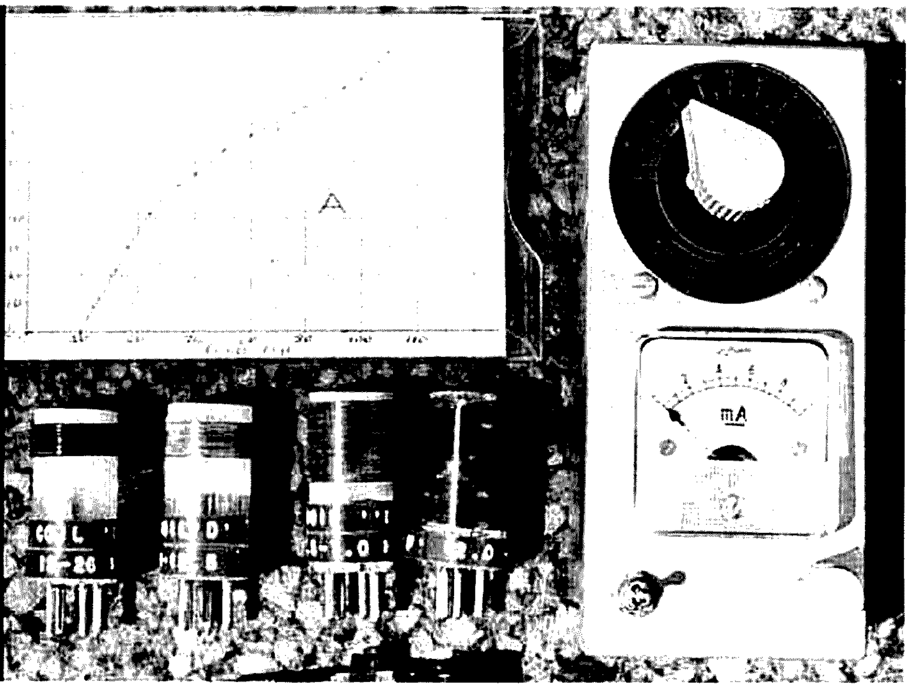
ham radio



frequency measuring
oscillator

Looking for a small, self powered gdo? This design offers a forward-reading meter and full coverage between 2-100 MHz

R. C. Alexander, W6IEL/7, Route 3, Box 502, Sequim, Washington

One of the most useful testing devices is the grid-dip oscillator. This instrument has evolved through many stages of development from tube-types to modern solid-state models known as gate-dip oscillators or tunnel dippers. One of the advantages of the later designs is user convenience – no need for primary power, which can be a problem when climbing a tower to make antenna adjustments.

frequency-measuring oscillator

Here's a still later development in solid-state gdos, but with a new (and more nearly correct) title and a different mode of indication. Instead of a dip in meter indication, the fmo meter swings forward with alacrity, which makes for easier observance of resonance.

The circuit is shown in fig. 1. Transistor Q1 is in a Colpitts oscillator circuit. The oscillator is followed by Q2, Q3; a high-gain dc amplifier of the Darlington configuration. Connections and polarities are such that meter movement is in the

forward direction when the circuit to be tested is near or at resonance.

construction

The fmo can be housed in a 2¼ x 2¼ x 5-inch utility box. If you desire this shirt-pocket size, the switch and pot should be of the miniature variety such as used in small transistor broadcast sets. For those wishing to duplicate the instrument shown, a parts list (**table 1**) and coil data (**fig. 2**) are provided. A full-scale layout drawing (**fig. 3**) is also shown.

A 2 x 3¼-inch circuit board holds the parts. Three of the circuit wires (shown dotted in **fig. 3**) are installed on the *underside* of the board to reduce capacitance effects. All other wiring is installed on top of the circuit board. Install the transistors upside down, with leads facing up, and bend the leads toward their respective soldering pins, which may be small brass push-through connectors or standoffs. Holes 1 and 2 in **fig. 3** are for mounting screws (see photo).

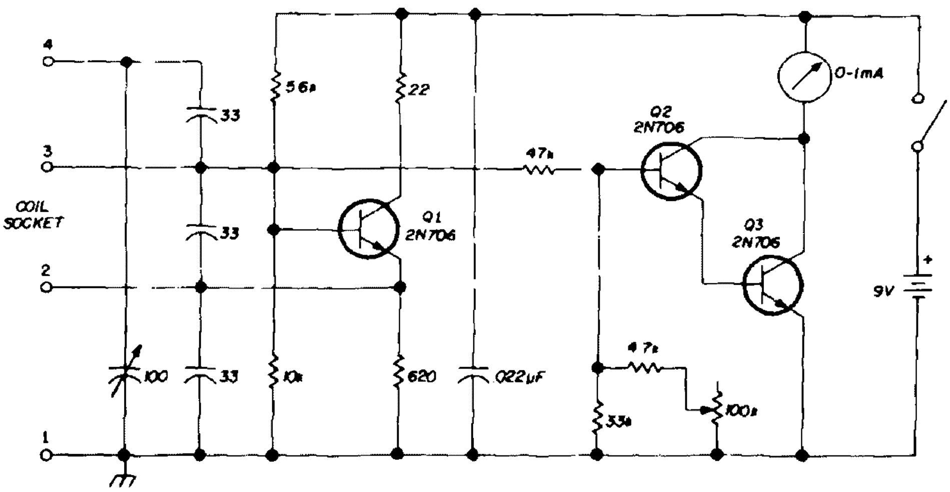


fig. 1. Schematic of the frequency-measuring oscillator. Circuit features forward-reading meter and full coverage from 2 to 100 MHz.

tuned circuit

The superior performance of the fmo at all frequencies is due primarily to the coil-capacitor arrangement, in which feedback and bandspread are obtained automatically. Except for coils A and B, each coil has a series-parallel capacitor arrange-

table 1. Parts list for the frequency-measuring oscillator.

qty	description
1	aluminum mini-box, 2¼ x 2¼ x 5"
1	2" 0-100 dial-plate
1	pointer knob, 1"; modify if needed
1	small knob for pot control
1	perfboard piece, 1/16 x 3¼ x 1-7/8"
2	3/8 x 1¼" mounting pillars, poly or steatite
1	miniature 0-1 dc milliameter, 1½" mounting-hole size
1	*miniature* pot, ½" diameter, 0-100k
1	miniature spst toggle switch
1	battery mounting strap
1	miniature 5-pin socket, Amphenol 78S-5S, with retainer ring. Modify per instructions
3	33-pF miniature glass or ceramic capacitors
1	.022-μF (or .02) mylar or paper miniature capacitors
1	0-100 pF variable capacitor, Hammarlund MAPC-100-B or equivalent, with ¼" shaft
7	¼ watt resistors, one each 22, 620, 4700, 5600, 10k, 33k, 47k
3	transistors, 2N706, npn silicon
7	miniature capacitors for coils, mylar or ceramic, three 100 pF, two 240 pF, two 470 pF
4	coil forms, Amphenol 24-5H, 5-pin. Miscellaneous wire, solder, and flea clips for terminals

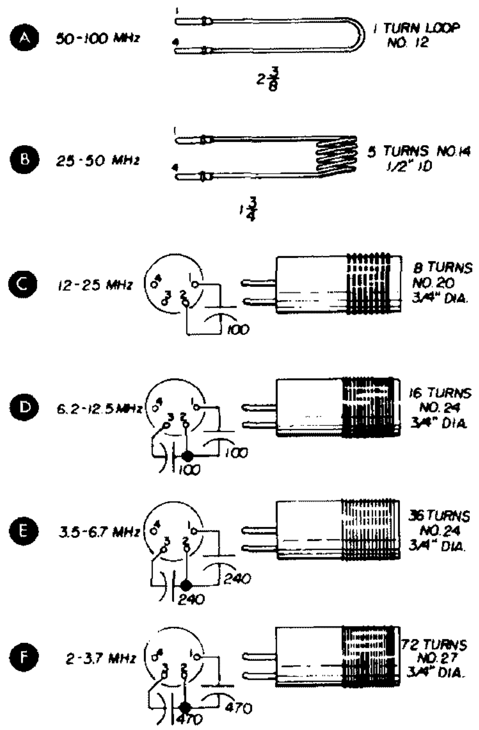


fig. 2. Coil-construction data. Pin 5 is removed from the four low-frequency coils and used for plug-in support of coils A and B.

ment, which provides the bandspread feature described.

Wind a few extra turns on coils E and F so wire can be removed if necessary to bring the circuit within the desired frequency range. This modification may be necessary as the small fixed capacitors will be within 10-20% of their marked values, and adjustment of the tuned circuit is made by removing coil turns. A light coat of Q dope should be applied to the coils. Final calibration (described below) is done after the coil dope is completely dry.

coils and sockets

The coil forms and sockets are modified by removing pin 5 of each form and the center shield of the socket. (These pin numbers are not used; the number sequence is shown in **fig. 2.**) The removed pins are used to support coils A and B in the socket.

When constructing the low-frequency coils, install the padding capacitors well down inside the coils, keeping the leads short. The lead from pin 4 goes to the top end of the coil; this is to minimize desoldering problems when removing turns from the top of the coil for calibration.

Keep all leads as short as possible

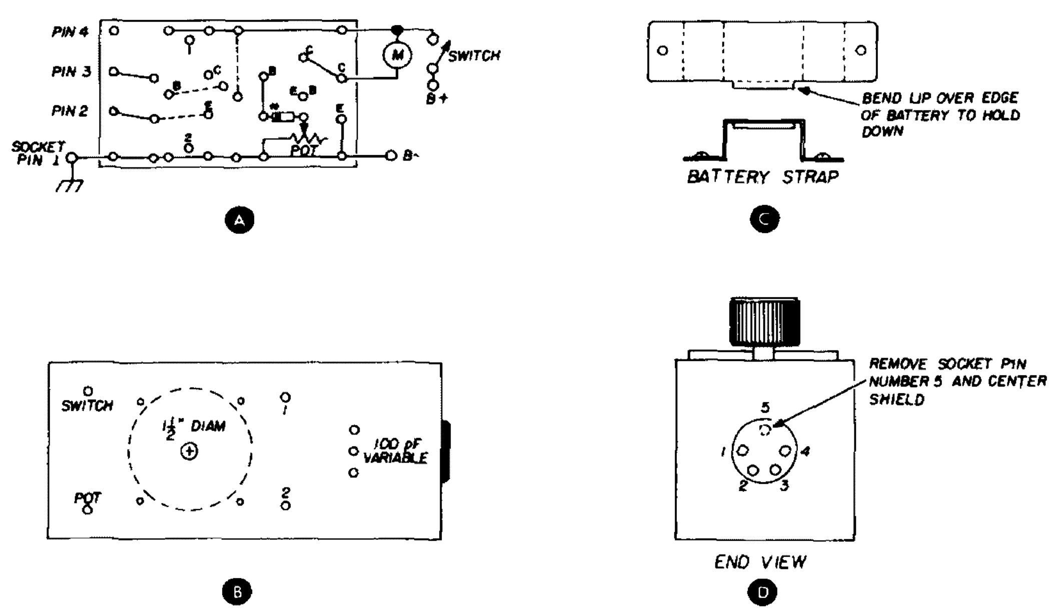


fig. 3. Full-scale layout drawings. (A) and (B) are templates for the circuit board and enclosure top. Holes E, B, C are for transistor leads. Component marked with asterisk is the 4.7k resistor in fig. 1, which was added after unit was built. Holes 1 and 2 are for mounting polystyrene pillars. Lower edge of knob (D) must be sawed off so set screw will reach capacitor shaft. Note that the numbers molded on the socket are not used. Three small holes in (B) are used to hold the dial in place.

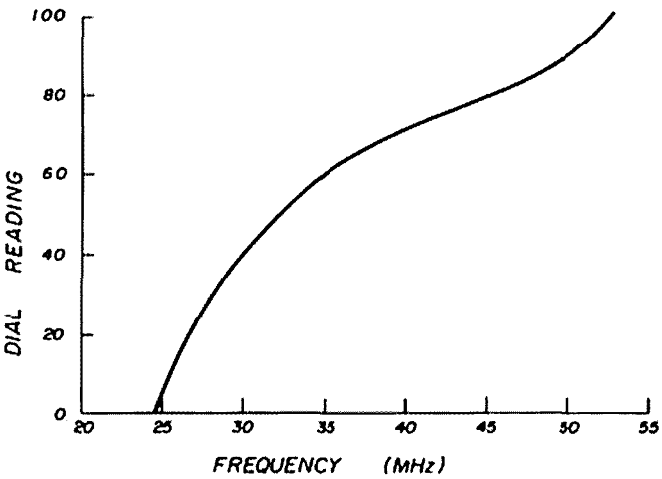
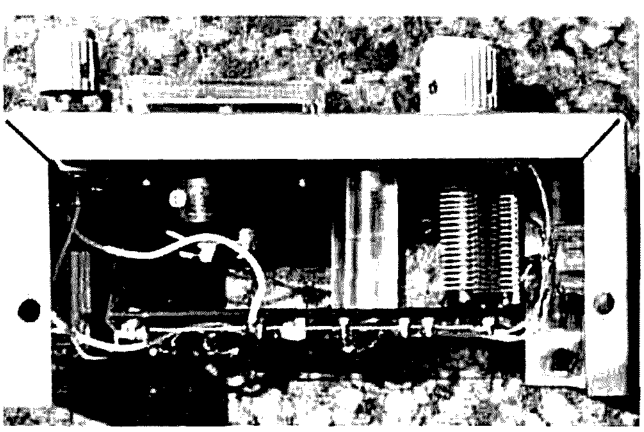


fig. 4. Typical calibration curve. Data is for coil B, 25-50 MHz.

between coil socket, variable capacitor, and the circuit board. Place a solder lug under *each* mounting screw for the variable capacitor, and run a wire from these lugs to pin 1 of the coil socket. Wire a short lead directly from pin 1 of the socket to the end of the ground wire (B−) along the edge of the circuit board nearest the coil socket.

Care should be used when soldering leads and capacitors into the coil forms as the material softens when excessive heat is applied. A heat sink, such as a pair of long-nose pliers with a rubber band



Side view of completed instrument. Objects to left of tuning capacitor are polystyrene support pillars.

around the handle, should be used when soldering heat-sensitive components. This also applies to transistors.

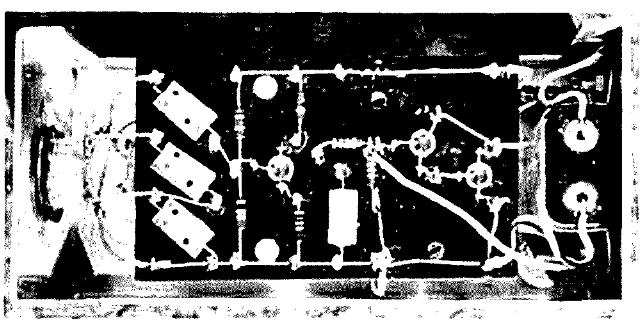
An inexpensive milliammeter with low damping action is suitable. A 0-1 mA meter is adequate; a more sensitive meter costs more and provides no added advantage.

calibration

The instrument should be calibrated against a receiver of known accuracy. Plot the calibration curves with at least 20 data points for each coil range. Fig. 4 is a calibration chart for the 25- to 50-MHz range. Accuracy will be only as good as the data source and the care used in calibration.

operation

The fmo is used in the same way as the earlier gdos; that is, when the circuit or



Wiring on component side of circuit board.

coil being measured and the fmo are tuned to or near the same frequency, the meter will swing forward. It will not dip as in a gdo. The pot is first adjusted to take the meter slightly off zero, the sensor coil is brought close to the coil or circuit under measurement, and the fmo knob is turned until a sharp forward meter indication is obtained. The fmo is then backed off. The smallest amount of coupling that will give a meter indication should be used.

The frequency-measuring oscillator is a versatile instrument. In addition to checking resonance of coils, antennas and other tuned circuits, it may be used as a CW signal source for circuit alignment. It's also useful for measuring inductance and capacitance within the range of component values used at rf.

ham radio

emitter-tuned preamplifier for 21 MHz

A simple and stable two-transistor project to improve receiver performance on the 21-MHz band

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Some of the commercial receiving gear I have owned has shown a pronounced need for more gain at 21 MHz. Measurements at my station show that the antenna noise at 21 MHz is 25 to 30 decibels less than at 7 MHz. The term antenna noise is used here to include all noise received by the antenna, whether it is atmospheric noise, galactic noise, man-made noise or whatever. Antenna noise is the increase in noise heard when the antenna is connected to the receiver. It is not intended that very strong local sources of man-made noise be included in this definition, but only those which appear randomly.

If the antenna noise cannot be heard, then many low-level signals cannot be heard either. Receiver gain should be sufficient to hear antenna noise, and a preamp connected between the antenna and the receiver is one way to bring the gain up.

In my case, a circuit was required which would be stable, have 25- to 30-decibels gain, bandwidth of about 300 kHz (I operate CW only), reasonable noise figure and adequate cross modulation and intermodulation performance.

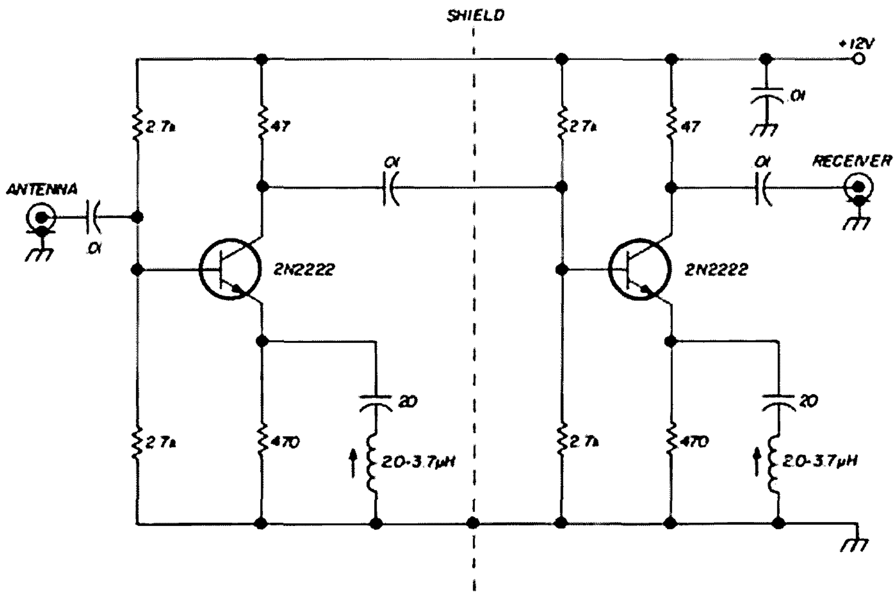


fig. 1. Schematic of the 21 MHz emitter-tuned preamplifier.

The first circuit I experimented with was a simple fet common-source amplifier, unneutralized, with parallel tuned circuits in the gate and drain leads. Stability was a severe problem, and the circuit was more likely to oscillate than amplify.

I decided on an emitter-tuned amplifier similar to one described by Chow and Paynter[1] would be worth a try. My version of the circuit is rather different from theirs because of the difference of intended application.

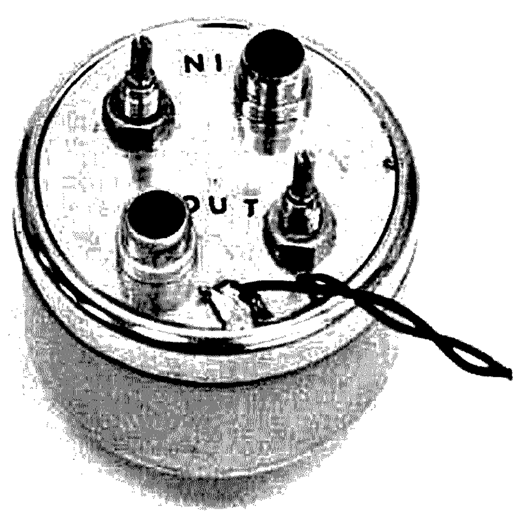
theory

The voltage gain of a common-emitter transistor amplifier stage is roughly equal to the load seen by the collector divided by the impedance in the emitter circuit. The intrinsic emitter resistance of the transistor must be included as part of the emitter circuit impedance when estimating gain. Its value is:

$$r_e = \frac{26}{I_E \text{ (ma)}}$$

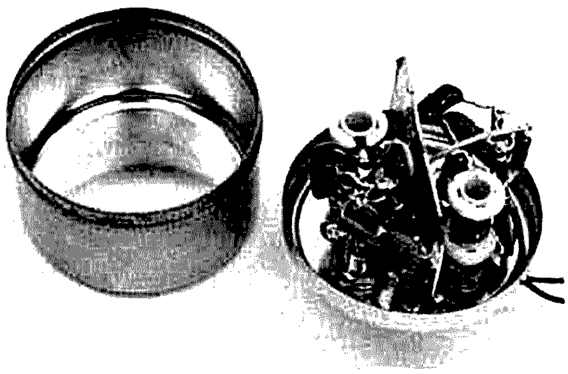
To keep r_e small so that gain will be high, each transistor is operated at about 10 mA emitter current; this results in r_e being about 2.6 ohms.

Fig. 1 is a schematic of the 21-MHz emitter-tuned preamp. Tuned amplification is achieved because the series-tuned emitter circuits present low emitter circuit impedances at resonance and high emitter circuit impedances outside the passband. The coils I used had an unloaded Q of about 120 at 21 MHz, and the 20-pF capacitors were duramica. Impedance of the series-tuned circuits at resonance should be about 3 ohms. Total emitter impedance, including intrinsic emitter resistance, should be about 5.6



Complete preamp is built in a can; hookup wire brings in power; phono jacks connect to receiver and antenna.

ohms. The collector resistor of each transistor is 47 ohms; if a 50-ohm load is paralleled with this resistor, the collector will see an equivalent load of about 25 ohms. Thus the voltage gain should be $25/5.6 \cong 4.5$ or about 13 dB. Two stages would then have a gain of 26 dB. Measured gain of the two-stage amplifier was 27 dB, indicating the above estimate was reasonably close.



All wiring is done on the can's lid without the use of tie points.

The 3-dB bandwidth is determined by the Q of the series-tuned circuits. Intrinsic emitter resistance must be included as part of the series-tuned circuit, but the 470-ohm resistor has negligible effect at resonance. With r_e included in the series-tuned circuit, Q is 66, giving a bandwidth of about 300 kHz. This was verified by measurement. Two stages decrease the bandwidth to about 200 kHz.

Input impedance to the base of each transistor was not measured, but since estimated gain and measured gain agree so closely, this impedance is assumed to be in the range of 50 to 100 ohms.

Outside the passband, the series-tuned circuits become large impedances, and the emitter resistance approaches a limit of 470 ohms, resulting in a gain-per-stage of 0.05 to −26 dB. This means −52 dB outside the passband for two stages. The floor of the response curve should be 79 dB below the peak amplification, but it is probably not this good due to stray leakage paths.

At first glance, this amplifier circuit may appear to have poor noise figure as well as poor cross-modulation and intermodulation performance. Measurements of this type were not made, however, no operational shortcomings were detected. The noise output of the receiver dropped 20 dB or more when the antenna was replaced with a 50-ohm resistor, indicating internal amplifier noise was well below the noise received by the antenna.

construction

The amplifier was built in a small steel can 2 inches in diameter by 1-3/8-inches high. All parts are mounted on the lid of the can as shown in the photograph. A brass shield, soldered to the lid, separates the two amplifier stages; holes are drilled through the shield to pass the interstage coupling lead and the 12 volt lead. Components are soldered together without tiepoints, and are supported by their leads. This results in compact construction and short leads. Hookup wire feeds in the power supply voltage, and phono connectors are used for input and output. The coil slugs should be adjusted for maximum receiver output at the center of the frequency range to be used.

conclusion

This amplifier is simple, easy to build and relatively fool-proof. Current drain is rather high, but stability is excellent. The emitter-tuned amplifier offers a quick and easy way to obtain rf or i-f gain. It may be used at other frequencies by using appropriate series-tuned circuits, but gain and bandwidth will be a function of Q. Less gain is to be expected at higher frequencies due to decreasing transistor h_{fe}.

reference

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ham radio

tuning
toroidal inductors

Professional tips on using simple test equipment and ingenuity to measure and tune toroidal inductors

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Toroidal inductors have **been praised** as the ultimate inductor for miniaturized equipment. Their usefulness extends from audio to vhf with power levels from sub-microwatt to several hundred watts. The magnetic field of the toroid is closed, and, for this reason, it is somewhat self-shielding and will operate well under crowded conditions.[1] This closed magnetic field also means that a grid-dip oscillator is going to be worthless because there is not enough flux leakage from the toroid to permit coupling. Put the grid-dipper on the shelf. It is time to develop some practical methods and procedures to tune those toroids.

equipment

You will need a signal source with a low distortion sine wave because it will give a better indication at resonance. A distorted waveform can be restored to a clean sine wave with a parallel tuned circuit to ground. Hewlett-Packard makes some beautiful signal generators but don't overlook the following possibilities:

1. Heathkit, Eico, and other budget-priced equipment.
2. Military and commercial surplus.
3. Homebrew oscillators.
4. The station transmitter (in spot mode or swamped).
5. A vfo.
6. Receiver audio (beat bfo against crystal calibrator).
7. Electronic organ.

A meter will be required mostly for use as a null detector. The possibilities for meters are a bit more limited but consider:

1. Any vtvm.
2. A diode and a sensitive dc meter.
3. A vom.
4. A receiver with an S meter.
5. An oscilloscope.

The problem of measuring frequency is a little more difficult. The dial calibration of some generators is adequate for some applications. More high-quality digital frequency meters are being found in ham shacks now for several reasons. Heathkit has a very reasonably priced counter kit, but possibly even more important to the ham is the switch that industry is making toward the new integrated-circuit counters. This has made available numerous low-frequency tube-type counters that still perform quite well. Counters are not the only

answer. Anything from a crystal calibrator to an electronic organ or a pitch pipe can be used to find the frequency.

equipment connections

Any of the five equipment setups may be used to tune any coil and capacitor. However, there are some preferences. It is generally best to tune the circuit as a parallel circuit if it is to be used in the equipment in parallel resonance, and best to tune in series if it is to be used in series resonance. An exception to this rule is at

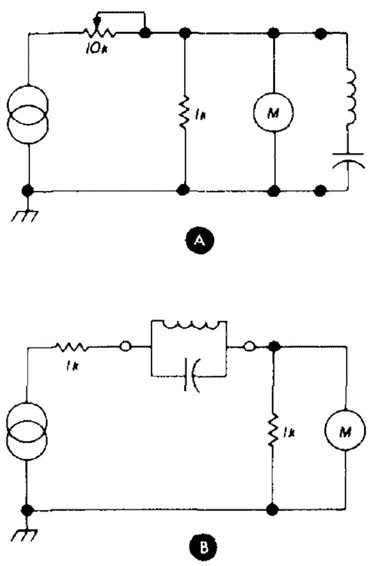

fig. 1. Test setups *for* determining *resonant* frequency of an L/C circuit using toroids. Both circuits will show a voltage minimum at resonance. A is series resonant and B is parallel resonant.

frequencies above about 2 MHz where it is desirable to tune *all* circuits in parallel to minimize the effects of lead lengths, even if the actual final circuit is series resonant.

Use a low signal level; ten millivolts is a good level, but unfortunately, most inexpensive vtvms will not indicate a level this low. If a good high-impedance vtvm is available but lacks the ten millivolt sensitivity, then **fig. 2** is the hookup to use with a volt or two at the peak reading.

It should be noted that the hookup of **fig. 2** indicates a peak or maximum voltage at resonance while **fig. 1A** and **fig. 1B** both indicate nulls or voltage minimums.

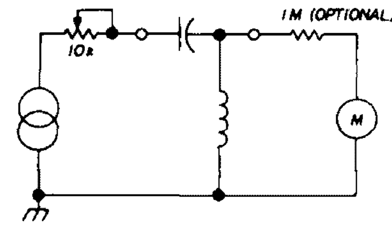

fig. 2. Setup for use with a high impedance vtvm lacking 10-mV sensitivity. The voltage will peak at resonance (usually 1 to 2 volts).

Fig. 3 requires an oscilloscope, but may be used with generators that have high harmonic distortion or noise in the output. This is a phase-shift method of tuning, and, if a good scope is available, this method is very accurate. Off-resonance frequencies cause an ellipse to be displayed which closes to a straight line at resonance. Distorted signals may be used, but the straight line at resonance will not be perfect.

Small coils and low-frequency coils require low test levels, but as physical size and frequency increase, the core is less likely to saturate, and higher test levels may be used. The cause and effect of saturation is too complex to detail here.[2] A good test to determine if the level is too high is to decrease the level to half or lower and check the resonance again. There should be no change in the resonant frequency. This test may be used to allow higher test levels on larger cores or at higher frequencies.

tuning to frequency

It is easy to get the inductance of a toroid in the ballpark by winding the turns determined by:

$$\text{turns} = 1000\sqrt{L_{mh}/A_l}$$

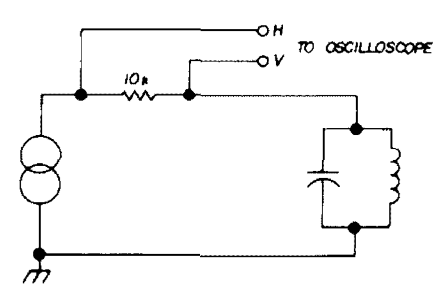

fig. 3. Accurate test setup uses an oscilloscope and the phase-shift principle. This setup can be used with signal generators with distortion and noise in their output.

L_{mh} = wanted inductance in millihenries.
A_l = millihenries per thousand turns on the core. (If A_l is given in microhenries per hundred simply move the decimal one place to the left.) A_l is sometimes stamped on the core but, most often, must be obtained from the manufacturer's data sheets.

Any one or a combination of the following methods may be used to fine tune the toroidal inductor:

1. Adding or subtracting turns. It is helpful to figure the hertz per turn or turns per hertz and compute the total turns to be adjusted. Raiding the wife's sewing basket for a crochet hook will accelerate turns removal. Obviously, if the coil has very few turns this method is not accurate enough.

2. All the remaining methods are variations of the variable capacitor method. The variable capacitor may be all or only a small portion of the total tuning capacity. Of course, true variable capacitors of the rotor/stator or piston type with dielectrics of air, ceramic, glass or quartz as well as the multi-plate compression mica types work very nicely; they are fine if you don't mind spending a mint just to tune a circuit.

3. The most economical variable capacitor on record is two pieces of magnet wire tightly twisted together and then trimmed to the desired capacitance — commonly known by oldtimers as a gimmick. Number 22 or 23 heavy insulated magnet wire is a good choice and yields from 1 to 5 pF per inch depending on how tightly it is twisted. Trim off a fraction of an inch at a time to raise the resonant frequency of the circuit; if too much is trimmed off simply twist the remaining wire a *little* tighter.

4. Disc-ceramic capacitors may be ground off with a small sanding disc or grinding wheel. Disc capacitors can be reduced in value by 50 percent with no deterioration in performance. Just be certain that the grinding action takes place longitudinally to the plates. In other words, don't smear any metal across the dielectric between the plates. Disc ceramic capacitors offer several advantages in this application. The entire circuit capacitance can be in the single disc capacitor. The disc can be a negative temperature coefficient type which

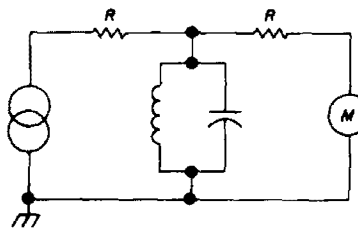

fig. 4. Circuit for determining Q. Make the two resistors as large as possible.

will compensate for temperature drift in the inductor. Large capacitances are physically very small in the low voltage, high temperature coefficient capacitors. Once the circuit is tuned the exposed capacitor edge should be coated with melted wax or polystyrene (Q-dope).

5. Spreading or compressing the turns of the coil will vary the turn-to-turn capacitance and change the resonant frequency slightly. This method is actually cheating and is mentioned only as a quick-cure method involving some risk. A toroidal inductor has a closed field only if it is perfectly symmetrical and any intentional distortion of the winding will cause flux leakage to the chassis, other inductors and other circuits. The best practice is to preserve symmetry. The start and finish of the winding upset the symmetry enough but are necessary evils.

Q measurement

Some of the equipment used to tune the toroidal inductors can provide information about the quality factor or Q of the coils.[3] Use the setup in **fig. 4** with nearly maximum output from the signal

generator and the largest possible values for the resistors. It is very important to not load the Q with a low-impedance generator or meter. If loading from the meter or generator cannot be avoided, it is still practical to make Q comparison tests to evaluate inductors. Adjust the frequency for a peak reading on the meter. Here again a low voltage level is preferred. Tune up in frequency until the voltage is 0.707 of the original reading. Record this frequency as F_h. Now tune the generator down in frequency until the level is again 0.707 of the peak voltage and record this frequency as F_l. These are the 3-dB down points, or half-power points. The formula that closely approximates Q from the 3-dB points except for Qs less than 10 is:

$$Q = \frac{\frac{1}{2}(F_h + F_l)}{(F_h + F_l)}$$

The tips in the preceding paragraphs are by no means limited to the tuning of toroids, but, since toroidal cores are not adjustable, some of these methods must be used. All of the methods are used daily in commercial practice, and are all practical. There have been so-called adjustable toroids available, but the adjustment was at only one point on the core which upset the symmetry and caused flux leakage and stray-coupling problems. It is also possible to grind notches in the core material of a coil with few turns and, if done at many different points around the core, a high degree of symmetry can be maintained. Core grinding is tedious and seldom used commercially.

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ham radio

nostalgia with a vengeance

A blonde, a kilowatt and a memory —

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It was during spring-cleaning of the attic under dire forebodings from my wife that it happened. For it was thusly, while investigating a pile of boxes situated in a long undisturbed corner, that I discovered the bottles. Antiquated bottles they were; mellow in their peaceful hibernation with thick coatings of dust.

Envisioning a sweet rendezvous with cherished brew of Ohmar's squeezings, I deftly brushed away the dust. Alas, the bottles were devoid of the coveted spirits! In fact, the bottles, though of ancient vintage, were quite empty except for the insolently protruding gizzards of primitive electron tubes. Closer inspection revealed enumerations such as 211E, 210 and 250.

Tempus fugit and shades of hades, how time flies! (In the manner of greased lightning, OM.) Shutting my eyes to black out the eerie illumination, sweet reminiscence conveyed me, dream-like, to days of yore. Ah, sweet days of yore. . .

In the wee small hours of a misty morning, I deposited the belle of the senior class on her doorstep and bestowed the usual prolonged kiss in the usual tender fashion. Then, not being overly-desirous of a QSO with her OM, I leaped into the waiting Essex.

With effervescent gusto I coaxed the last rpm from the protesting connecting rods. Soon, I was home with my first and true love, the 80-meter Hartley with one 211E, three 210s and two 250s coordinating more-or-less in parallel.

I reached up and yanked the handle of the main switch. Fondly, I beheld the glowing filaments. (Brother, you too would glow with double voltage on your filaments.) Trembling with anticipation of a hot DX contact, I appraised the performance of the rig in the usual professional manner.

I pressed the key for a few brief moments. A 500-cycle roar of defiance shattered the ether. The motor-generator tugged in agony at the half-inch bolts which secured it to the four-by-four support, then resigned to its torment by dropping speed. The lights in my room dimmed by an amount deemed proper.

I picked up a pair of binoculars and observed the light bulb in my neighbor's garage, some fifty yards due South. The intensity of its glow indicated healthy antenna radiation. I released the key just as the filament of one of the 250's became visible through the plate. Feeling

well rewarded for my efforts, I *turned* my attention to the receiver.

The receiver, though completed only ten days previously and not completely de-bugged, had already revealed itself as a signal snatcher *sui generis*. It was, in fact, the reception of a VK from Australia that motivated the DX quest about to take place.

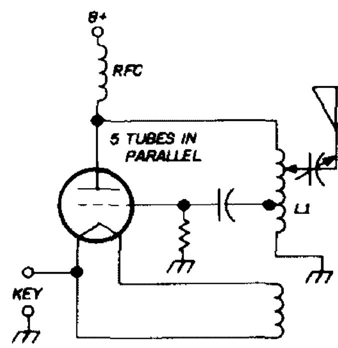


fig. 1. Salient features of the transmitter. Abundant harmonic generation assured a response on some band. L1 is copper-tubing tank from a Hoyt automatic water heater. Plate voltage was 1200V on spaces, 1000V on dots and 850V on dashes.

The design, my own, consisted of a regenerative rf stage, regen detector in a reflex arrangement which re-squirted the signal back into the rf stage, which now functioned as an audio amplifier. This was followed by two stages of conventional audio amplification, or *almost* so.

The slight departure from the "run of the mill" involved the use of Model T spark coils as inter-stage transformers. The tubes, chosen after much deliberation were 201As.

I placed the 8000-ohm Baldwin phones over my noggin and turned on the receiver. It was, verily, full with the joy of life. As I advanced the various controls, a terrific fringe howl sallied forth from the cans and reverberated within my tortured head like a freight train stalled in a tunnel! Quickly, I jerked the phones off and backed down the regen controls, lest the OM burst into the shack with an R9 lecture about the need of sleep for growing boys, etc. Swearing with R9 fervor, albeit in hushed breath, I resolved to be more careful with those very, very *critical* knobs.

Using the utmost caution to maintain regeneration just on the threshold of oscillation, I searched the endless kilocycles for a voice lost in the wilderness. It wasn't long in forthcoming. Somewhere in the depths of Boltzman's constant, all but buried in the crackle of shot noise, shot capacitors, themal agitation and ionized soldering paste, I heard it . . . yes I heard it, a weak CQ!

Carefully, ever so carefully, I advanced the you-know-which controls. Now, a tiny bit more, a shadow here, an Angstrom there. The results were no less than magical. The signal, except when fading obliterated it entirely, now was R3 in any ham's lingo. Only thing was, the fading was somehow mysteriously synchronized with the announcement of the station's identity.

Carefully, ever so carefully, I trimmed the other controls with my left hand, while my right maintained vigilance with the detector regen control. Waiting in suspended animation for the final station identification, I arose from my chair. I leaned over the receiver to provide the nth degree of trimming by means of body capacity. A bead of perspiration dripped from my brow and landed dangerously close to the grid terminal of the detector tube. I blew the saline droplet away from the socket.

Then it happened . . . I caught the call just before he signed off. It was none other than the VK recently logged! Shades of hades, could I work him?

I laid the earphones on the operating table and promptly went to work with the home-spun bug. This was constructed from materials distinctly out of the ordinary, among which were the innards of a Pocket Ben, ignition breaker points and a segment of corset stave donated by the belle of the senior class. Between the bug and raucous note, I was certain my signals had that spark of personality needed to attract the attention of DX ops.

For the first minute or so, my sending fist trembled and threatened to freeze. However, the bug gave unselfishly of its

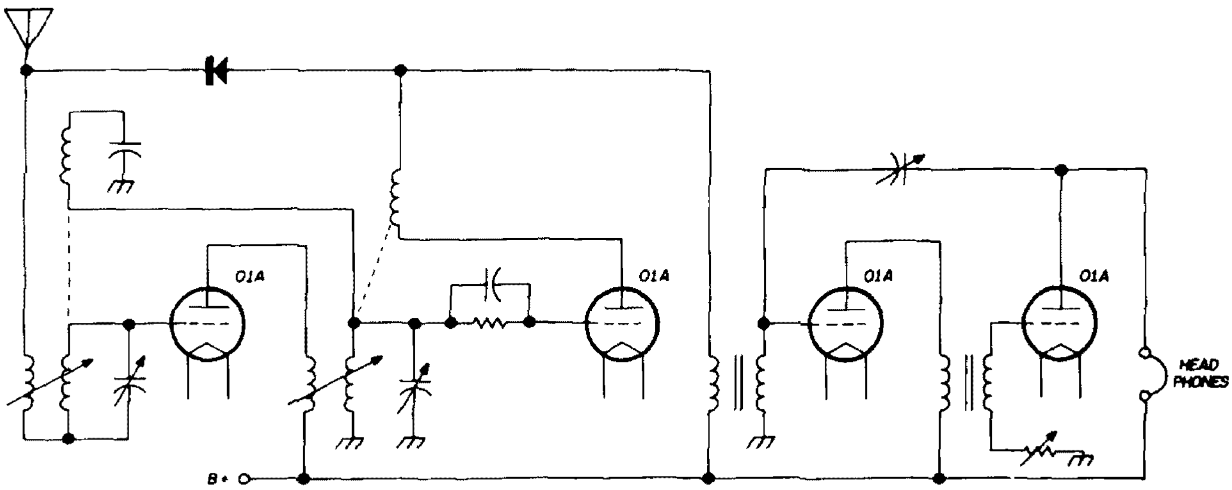


fig. 2. Schematic depicting general idea of the receiver. This set is regenerative to the nth degree; it should not be confused with workable schemes.

dots and I soon regained poise. I settled down to a fifteen minute grind of three VK calls interspaced by two of my own.

The dit-dahs were now oozing with the solid self confidence that I pounded into them. My enthusiasm mounted until I could scarcely contain myself. Finally, I repeated my call eleven times and culminated with a stately 'RK'!

I climbed into the phones and listened. Amidst the playful electrons, I heard *nothing*, absolutely *nothing*. Ever so carefully, I applied torque, one part physical, nine parts pyschic, to the detector regeneration knob. The impending fringe howl was lurking in the infinitesimal depths of perception, intent upon springing out at me, tiger-like. Precariously but nonetheless skillfully, I held it at bay.

Within the innermost realm of consciouness, I became aware of minute fluctuations in the background noise having the characteristics of dots and dashes. I copied it mentally: "sure gld to wrk the West coast OM. Ur sig FB here in Melbourne. Pse give my report . . . " The signal faided out but I could sense he was signing over.

I prepared to acknowledge his message. So intoxicated was I with the prospect of bettering my previous DX record by some seven thousand miles that I had to deliberately pause a few seconds before tickling the bug.

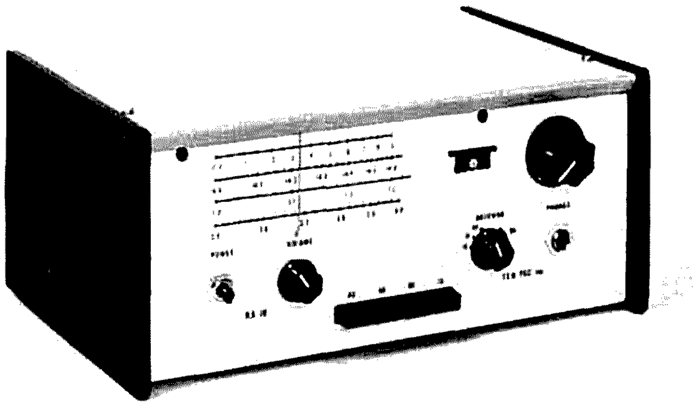
Then, *boom,* like a bolt of lightning did the terrible catastrophe strike! The phones, still on my head, became the forerunners of the present day dynamic speaker. Unmerciful decibels assailed by burning ears. Between the jack-hammer thumps, the fringe howl lashed out with tongues of sonic flame.

I tore the phones from my head and threw them on the table. The thumps were saying, "VK3– de W6– tnx for fb rept. Ur coming in QSA 5, R6." "R6! Shades of hades," I muttered to myself, "All he has is a superhet with no regeneration and he has the gall to call a signal R6 which I can just barely pick up."

W6– resided in the high-falutin section of town because his OM was well heeled, the owner of a chain of shoe stores. W6 attended private school, owned a brand new Model A Ford and had a 5-kW rig full of store-bought components. What with all this, he now runs away with my DX station!

So what? Did I begrudge his monied pop? Was I envious of his classy rig? Did I resent his swiping my VK? Shades of hades, *no*. After all, it was I who was the steady of the belle of the senior class! But OM, that is another story altogether . . . anyhow she's mighty anxious that I get this attic cleaned up, pronto!

ham radio



improved selectivity for direct-conversion receivers

The subject of these mods is the Ten-Tec RX10 — however, they may be applied to any direct-conversion receiver with good results

Vladimir N. Gercke, K6BIJ

About fifty years have elapsed since radio amateurs replaced their one- and two-tube regenerative receivers with the superheterodyne circuit. Progress has been spectacular . . . Or has it?

It is now possible to build a receiver for the high-frequency bands that performs as well as or better than the superhet without the problems caused by i-f amplifiers, such as images and spurious oscillations. The improved circuit is based on the old synchrodyne detection method and is known as direct conversion. It has been described in earlier issues of *ham radio*.[1, 2] A direct-conversion receiver available commercially is the RX10.* Reviews of the RX10 have appeared in *ham radio* and elsewhere.[3, 4]

To improve the selectivity of my RX10 I developed the front-end modifications described here, which are adaptable to other direct-conversion circuits as well.

Selectivity is provided by a sharp audio filter preceding the audio amplifier and the high Q of the input tank circuit. The high Q is developed by regeneration; when regeneration is removed selectivity will be about equal to that of a galena crystal set.

*Manufactured by Ten-Tec, Inc., Sevierville, Tenn. 37862.

Fig. 1 shows five diode-detector circuits and their respective selectivity curves:

A. Crystal or tube detector with no tuned circuit.

E. Same as D, but with extremely low L/C ratio.

As shown in **fig. 1**, the culprit is dimension A, which is the portion of the

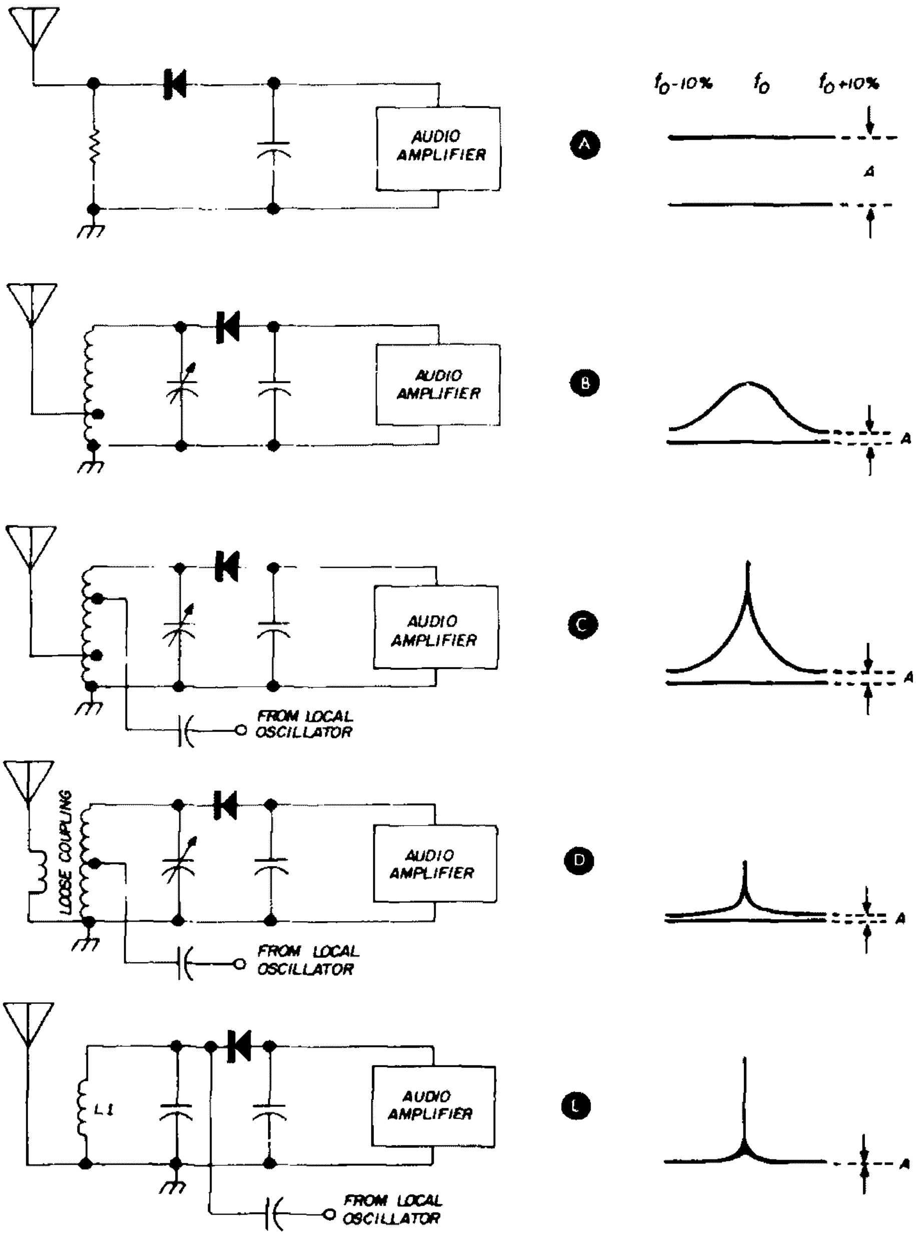


fig. 1. Five diode-detector circuits and their respective selectivity curves. Circuits C, D and E are direct-conversion types. Bias and audio filters have been omitted for clarity.

B. Same as A with added tank circuit.

C. Same as B with injected local-oscillator voltage.

D. Same as C with very loose coupling to the antenna.

response curve that levels off and continues indefinitely above and below f_0, the frequency of interest.

Loose coupling, as in **fig. 1D**, will decrease dimension A almost to zero; but most of the signal is lost. An audio

system with extremely high gain will be needed to restore the signal.

RX10 modifications

The circuit of **fig. 1E** seemed to be the solution to the selectivity problem. The which can otherwise defeat the purpose of the circuit.

All my experiments were made with the Ten-Tec RX10 receiver. The front end was modified according to **fig. 2,** a partial schematic of the unit. The value of

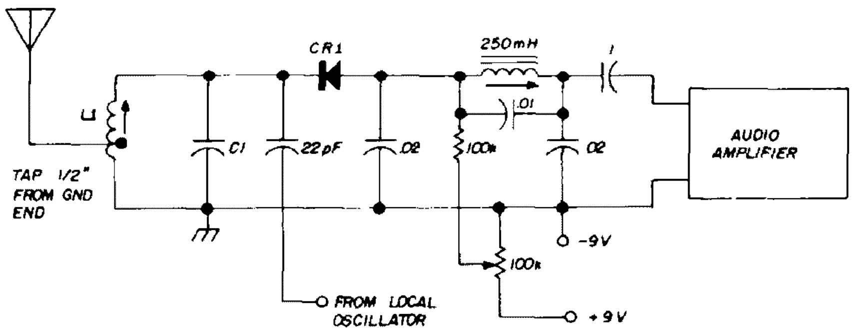
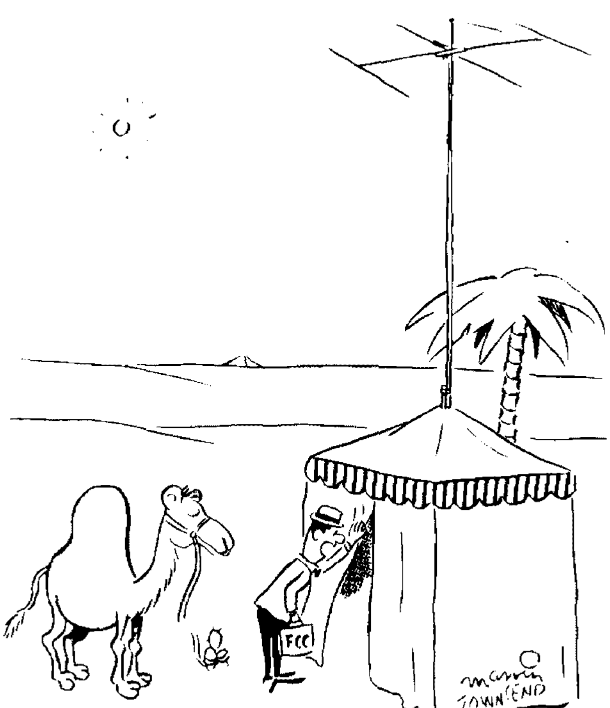


fig. 2. Modified front end of the RX-10. L1 is 2 turns no. 12 solid copper wire, 5/8-inch diameter; C1 is 1400 pF, 4400 pF and 13500 pF for 14, 7 and 3.8 MHz, respectively. CR1 is any hf or vhf diode or a transistor E-B or C-B junction chosen for minimum noise (varies to 10 dB among units with the same 1N or 2N prefix.)

tank coil was made with two turns of heavy wire, and the tap for the antenna was ½ inch from the ground end. All signals below f_0 are returned to ground, and all signals above f_0 are blocked by the enormous value of C1 (13,500 pF for 80 meters). C1 consists of several silver micas in parallel to reduce lead inductance,



"Sir, are you aware that you are interfering with television reception in this area?"

L1 is a compromise; it's too large for 20 meters and too small for 80 meters. One turn for L1 and 2,400 pF for C1 would be better for the 20-meter band.

conclusion

The selectivity of a tuned circuit is a function of its Q, the ratio of reactance to resistance at the resonant frequency. The excellent selectivity of the modified RX10 front end is due to the high Q obtained by decreasing the L/C ratio and by decreasing circuit resistance with large-diameter wire in the tank coil.

Perhaps we've been on the wrong track by climbing the vertical wall of the superhet for 50 years. Maybe we should go back and see if there's a path, or freeway, around that wall.

references

1. Richard S. Taylor, W1DAX, "CW Transceiver Operation with Transmit-Receiver Offset," *ham radio,* September, 1970, p. 56.
2. Rick Littlefield, K1BQT/6, "The Sideband Minituner," *ham radio,* October, 1970, p. 16.
3. Skip Tenney, W1NLB, "Ten-Tec RX10 Communications Receiver," *ham radio,* July, 1971, p. 63.
4. Morgan W. Godwin, W4WFL/1, "Ten-Tec Model RX10 Communications Receiver," *QST,* July, 1971, p. 44.

ham radio

digital station accessory

Concluding construction details, including the input amplifier, and counting circuits for direct frequency read-out of your receiver or transmitter

E. H. Conklin, K6KA, Box 1, La Canada, California 91011

My first counter, fortunately, used the relatively slow Fairchild DTL 9093 dual flip-flops. Two of these and a gate can divide by ten, with no input circuitry, from low audio rates up to 2.2 MHz. For many digital dial read-out purposes, however, these will not do for the first decade or so. Higher-speed TTL units are comparable in price, but take more care in driving the first FF to make them toggle (and produce a divided output).

Later, the decade dividers became available at reasonable prices. The ones tried here – the Sylvania SM90 which has no counter outputs and the 7490 made by Texas Instruments, National Semiconductor, Philco, Motorola and others – all toggled directly at low audio rates as well as at rf. In normal use these are up-counters; although with decimal read-outs like the Nixie, down-counting can be arranged by inverting the wiring.

Data sheets indicate a maximum frequency of 10 MHz, minimum, and 18 MHz, typical, for the 7490. My tests showed that some ICs toggled to a maximum of 30 to 41 MHz without an input amplifier, and much higher with certain amplifier circuits. Care must be taken to bypass the plus V_{cc} to prevent spikes from causing a miscount which can occur when noise or hum is present. Some types of input circuitry require relatively high input at low audio frequencies. Careful design of input circuitry for proper wave shape and signal level may be required to toggle the ICs from low audio frequencies up to extreme frequencies.

For most purposes a selected SN7490N decade can start the count string, following a satisfactory count gate. Tests on the low-cost SN7400N quadruple two-input gates indicated that a few operated above 110 MHz. One high-speed *SN74H00N went to 45 MHz.*

The JK flip-flops, such as the Sylvania SF7473N, were somewhat variable when tested on audio and rf. Several went to 33

or 34 MHz, but others did not. Therefore, if one of these units is used in the first counting position, selection is desirable.

Fully programmable up/down counters can be preset readily to start at any number so that division can be by any number. Most interest has been in decade operation, such as is provided by the SN74192N synchronous, programmable up/down decade counter. Input clock frequency typically can be 32 MHz. Power dissipation is greater than the 7490 decade counter.

The 74192, when used as the initial count FF, requires a count gate. Note that down-counters may be needed in only some decades. That is, if a receiver dial reads only to 0.1 kHz, and a one-second counting gate is used, the units and tens decade counters can be 7490 up-counters because they do not affect the read-out. Furthermore, the MHz counter also can use 7490 up-counters if they do not affect the read-out.

fier can be very strict if the digital station accessory is to be used for many purposes. It should be possible to obtain suitable drive for the first count IC, U16, with any input from about a tenth of a volt, to 25 volts or more. The impedance should be high so it does not significantly disturb the circuit to which it is attached. It should require no tuning or adjustment; it should respond to frequencies from below five hertz up to 50 MHz or so. Also, it should drive the first count IC reliably.

Although the 7490 decades will respond over such a frequency range, it is not always convenient to connect the IC directly to the circuit. This has resulted in a great deal of experimentation, and some of the more difficult problems were solved while resting in the middle of the night!

The basic circuit started with the seven-transistor amplifier used in the new 80-MHz Heath SM-105A counter. This

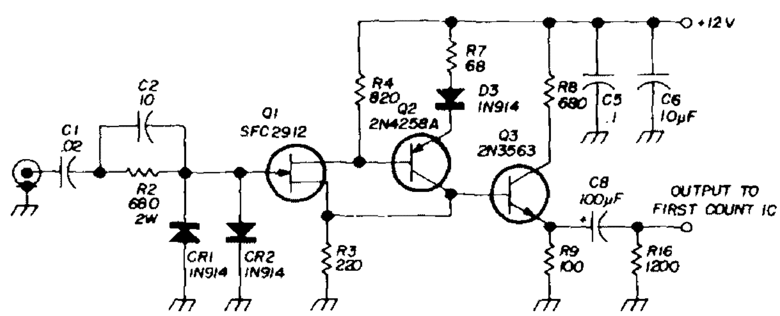


fig. 1. Input amplifier. See text regarding bypass of R9 and variations in resistor values. Q1 is Heath (TI) 417-251 at $2.25; Q2 is Heath 417-260, 80c; Q3 is Heath 417-125, 65c. Noise generation somewhere ahead of Q1's input gate was minimized by connecting 30 pF from final output to ground.

Dc wired information from a bandswitch can be provided to the megahertz read-outs without any counter, latch or decoder/driver for MHz information. Of course, this works best on amateur band equipment, but MHz read-out probably would be eliminated on all-wave receivers where 30 switch positions may not be tolerated anyhow. To clarify this, note that the down-count is a frequency division just like an up-count. The only difference is the direction of the BCD count going to the read-out display.

input amplifier

The requirements for the input ampli-

circuit was found to be sensitive as to selection of voltage, components and part values. The substitution of the 2N5485/MPS107 jfet, an inexpensive plastic device, proved to be subject to problems. Some of these problems did not affect another amplifier of similar design which used the metal-cased SFC2912 (Heath 417-251, $2.25), the case of which can be grounded. Somewhat improved performance may be obtained by further variation of resistor values from those shown in fig. 1, but the values shown do work.

Using a large output-coupling capacitor, and a variable sink resistor across the

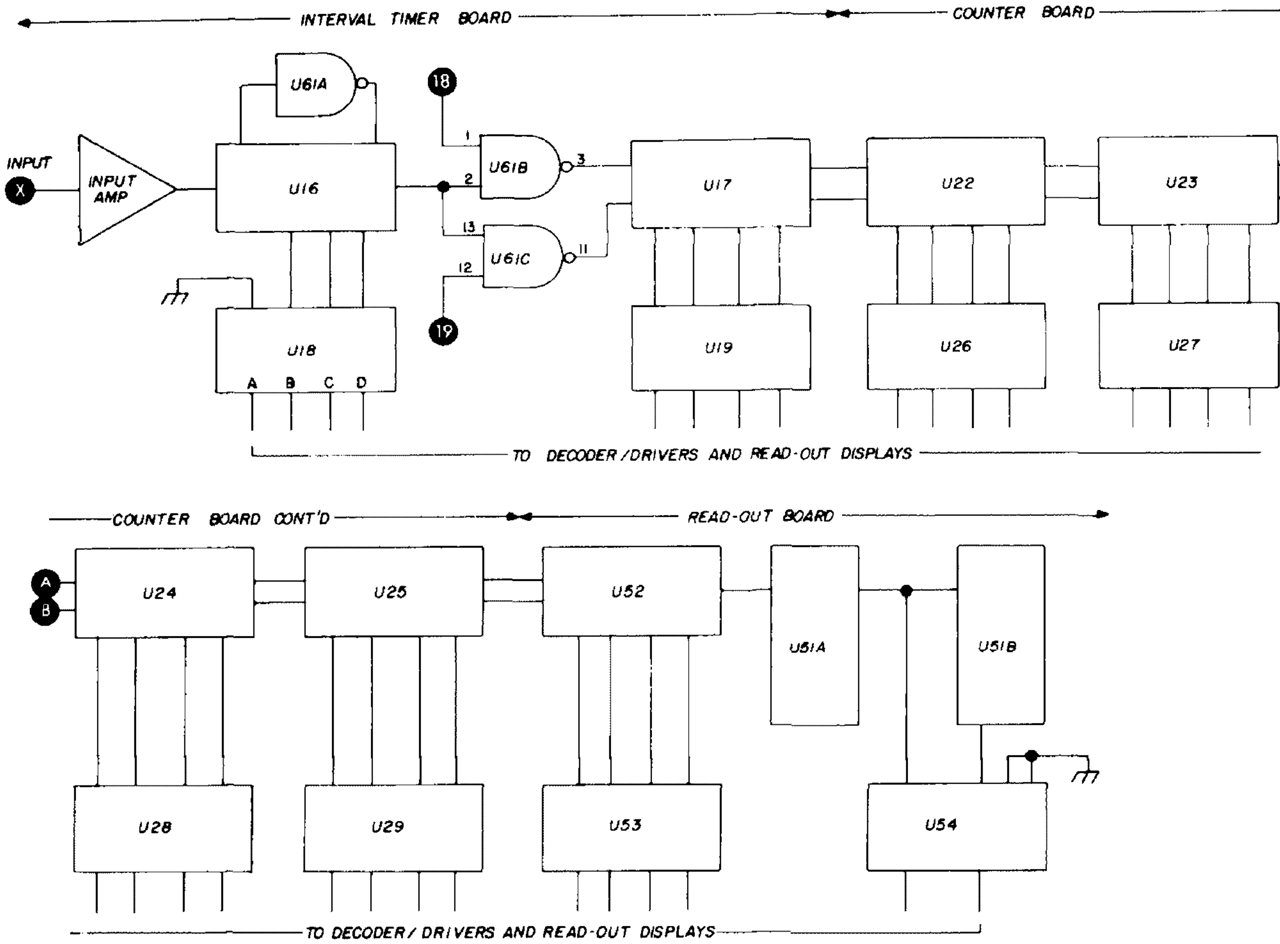


fig. 2. Flow chart for counter ICs. Circle are plugs.

following IC, this amplifier successfully drove a selected 7490 FF to 65 MHz. When the capacitor and a fixed sink resistor were installed permanently with shortened leads, there was some sacrifice in performance, but results were still adequate. The first IC, the A section of a TC7490E,* followed by a 74H00 gate, toggled accurately to 52 MHz when R9 in the amplifier was by-passed with several microfarads. A Philco 7490 counted to 34 MHz.

A 15-μF capacitor temporarily installed across R9 increased amplifier output and produced a square wave. However, when the leads were shortened and the fixed resistor substituted, the capacitor was omitted to prevent possible injection of noise which might result in miscounting; later, I found that toggling of the first IC on audio frequencies was being affected by spikes on the 5-volt power supply line. No further attempt was made to readjust the value of R16, the IC input sink resistor, for maximum frequency; counting from below 4 hertz up to nearly 50 MHz seemed adequate.

The input amplifier occupies about one square inch of the upper rear corner of the IT-board, and does not interfere with the mounting of the DIP ICs on that portion of the board which carries the V_{cc} and ground busses. Also, it leaves space for the end-mounting of steering diodes for multiple programs for the up/down counters. The top edge buss had been opened for use as a plus 12-volt V_{cc} buss for the input amplifier (connected to Plug-Y). The remainder of the edge buss remains at ground potential, plug-Z.

count chain

Vector sockets were provided for the count gate, U61A, and the first two count ICs, U16 and U17, shown in the

*Available from Solid State Sales, Box 74, Somerville, Massachusetts 02143.

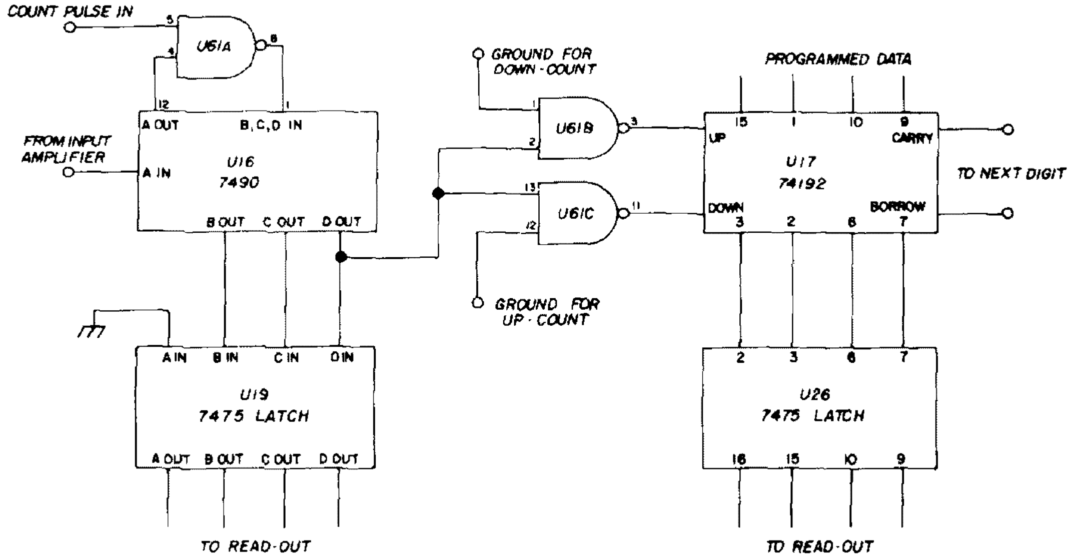


fig. 3. Flow chart for input gates and first two counter decades.

block diagram, **fig. 2**. These facilitate testing individual ICs and selecting the most satisfactory one for these critical positions.

Ordinarily, the input signal passes first through a count gate which controls the time during which the count is made, and the output drives the first FF. However, this resulted in some limitations (later found to be associated with noise spikes on the power supply) which were readily overcome by an unusual course. The A section of a selected 7490 decade up-counting divider, U16, was driven directly by the amplifier; this was followed by the count gate U61A which drove the BCD section of the 7490 (see **fig. 3**). This odd arrangement allows the A section to continue to count; the effect on the read-out is eliminated by tying the U18 latch input to V_{cc} or ground, whichever tends to correct the normal one-count error due to the relative phase of the time-base gating and input wave. At most, it will produce an error on one hertz in the count.

If part of the same gate is used for coincidence-gating, such as was done in U61D, there will be some interaction in the outputs of all gates in the same IC, even when V_{cc} is adequately by-passed to ground. Although this possibility should be kept in mind it did not cause any trouble in my unit.

Following the A section of U16, up and down gates U61B and C were installed for later use with the 74192 up/down counters. If 7490s are used these gates should be omitted.

The 7490 up-counting decade FFs may be used for U17, U22, U23, U24 and U25. These will be satisfactory for normal counting functions and for counting the synthesized operating frequency.[1,2,3]

The outputs of the counting decades are connected to the numbered D inputs of their respective 7475 memory latches, U20, U21, U26 through U29, U53 and U54. For use with MSD047 or SN7447N decoder/drivers and Minitron read-outs, the latch Q outputs are wired to the decoder/drivers on the read-out board.

Note that the V_{cc} switching scheme permits wiring the count latches on the IC-board to the AND gates which control clock information (but capacitors must be installed across V_{cc} switching contacts to keep the clock from advancing). Note that the final two decades are on the read-out board. The last of these is a dual JK FF which is adequate for a count of four; it repeats if the frequency is 40 MHz or above.

The general wiring of the 7490 ICs is much the same as described last month, except that it is necessary to reset all of the count decades after the end of the

count and after the transfer pulse, to display the count, and be ready for the next period. This means that there never is a grounded R_0 reset input in the count

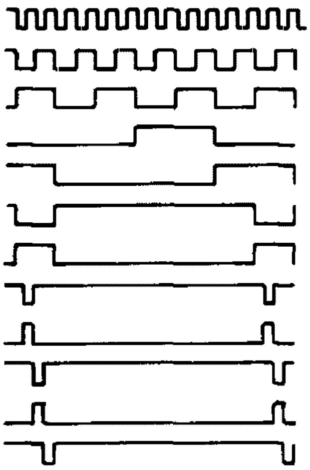


fig. 4. Pulses from resolution input, through divider U20 and U21, and selection of coincidence gating to generate the necessary pulses.

chain. However, one R9 input on each IC should be grounded.

gating

The gating system follows the general plan described by Kenneth Macleish, W7TX.[2] However, there are many ways to accomplish this.

By charting the divide-by-two and divide-by-three square waves at the end of the time-base chain, **fig. 4**, it is easy to select gating inputs that will provide a one-second count period (or decimal fraction). After this, there must be a short off-period containing a transfer (memory latch) pulse and a reset pulse. For 74192 up/down counters, there also must be a load pulse which transfers the programmed frequency correction into a presetting of the 74192 counters from information at their data inputs. These pulses can be produced simply by coincidence gating. Connecting the gates to the correct square-wave output produces the desired pulses at the right times.

Although this method is widely used there may be a small error in having reset pulses begin or end with the finish or start of the count period. This is because the propagation time through an IC may not be the same as the time required for resetting it. Therefore, the positioning of these pulses in the noncounting period has been selected differently.

Also, the divide-by-twelve SN7493N, used by W7TX as a final time-base divider, was replaced by two 7473 dual JK flip-flops. One of these could have been the 16-pin SN7476N, which is available at surplus prices. This has a common clock for the two FFs so, generally, it may be used where that is satisfactory, including the divide-by-three circuits in the time and clock sections. The 7476 is provided with separate clear and preset pins which are useful for counting days and months which start with "1".

Fig. 4 shows the pulses in the final time-base divider, using the two 7473s. The entire gating schematic is in **fig. 5**. Note that the Q and not-Q outputs are useful to avoid the requirement of a reversing NAND gate.

The load function in the 74192 up/down counters must be eliminated, or the load data be zero, in every case for unprogrammed counting of some unknown frequency. This can be done in several ways, including running the load signal out to a switch and back again. **Fig. 7** includes gate U60D, otherwise unused, which can be used with another NAND gate (or both can be replaced with one AND gate) to give one-wire control over a program loaded into the up/down counters.

Some resets, such as that for the 7473s, go to ground to reset. Others, like the 7490 and 74102, must go to a logic high. Because there is a fan-out limit of eight to ten in DTL and TTL gates, the reset signal for the 7473s is wired out to other boards on plug-D. NAND gates U62C, U70C and U80A invert the reset signal on those boards as necessary for 7490 and 74192 ICs.

Each memory latch presents a load of four FFs to the transfer pulse. Thus, only two 7475s are being driven by one gate section. An inverted transfer pulse is wired out to other boards, then inverted by NAND gates U62D, U70D and U80B.

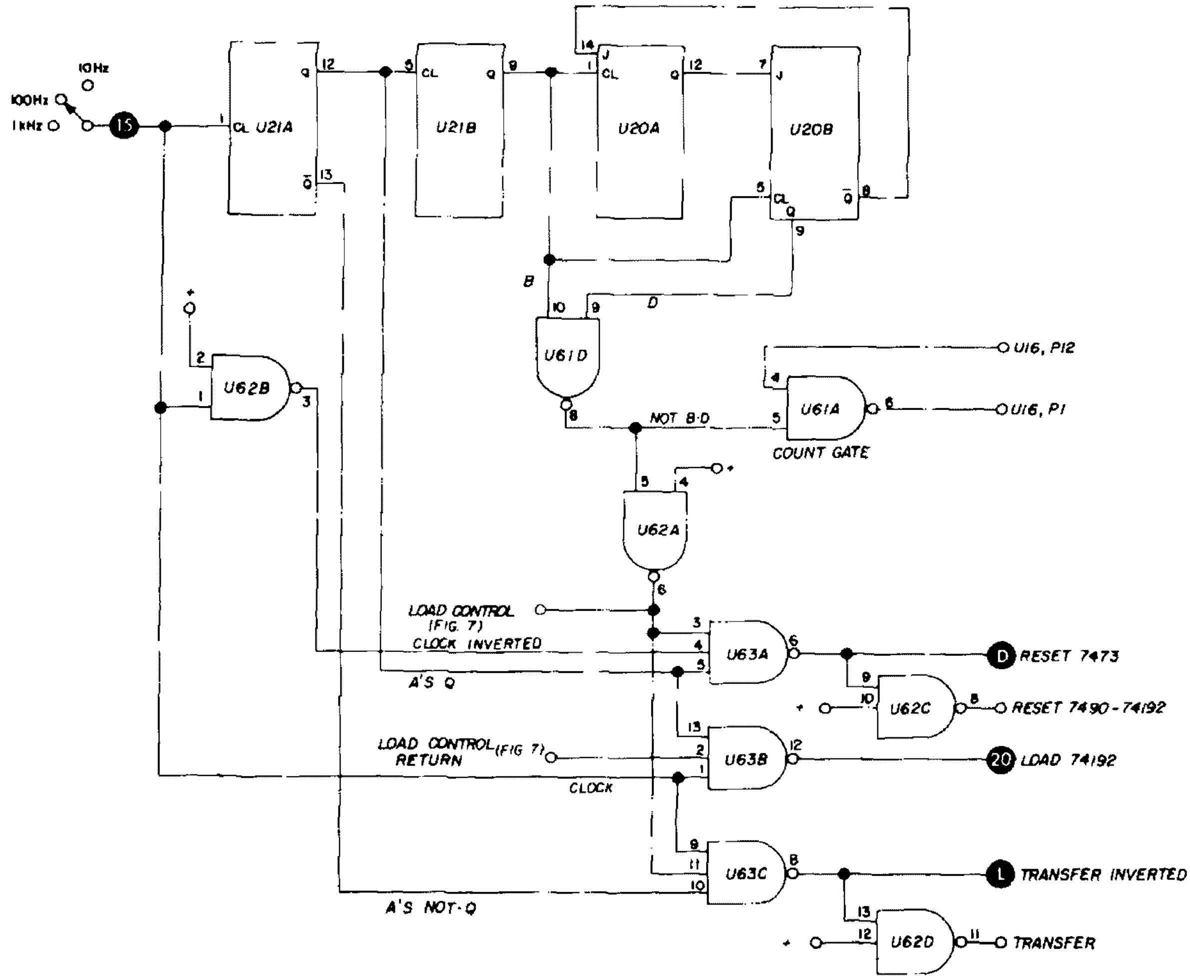


fig. 5. Detailed flow chart from resolution to generation of pulses, with pin designations. See fig. 7 for single-line load control using two NAND gates or one AND gate.

Three "resolutions" are provided at plugs 12, 13 and 14. The one-second gating counts units of hertz; the one-tenth-second gating counts tens of hertz; and the one-hundredth-second period produces hundreds of hertz. For some fast-tuning applications the shorter count periods are useful. Also, during CW and RTTY keying, the short periods will occasionally produce a stable read-out, while others will be short-counted due to keying during the gate-open time. This occurs in the "signal mode" of the synthesis method, but not in the vfo count method that will be described later.

When up/down counting is used with the loading of a correction, it would be complicated to change the programmed loading for more than one counting period. Therefore, only one resolution is suggested: Whenever the switches inject a programmed correction which is satisfactory to the operator.

The different pulses can be seen on a scope but it is somewhat of a problem to identify where each one occurs in the non-counting period. In case of trouble, try disconnecting one source of these pulses and study what happens. When the coincidence gating plan is worked out and followed, however, everything turns out right. Nevertheless, keep in mind that a reset pulse before a transfer pulse will give a read-out of exactly nothing. A reset pulse after a load pulse in the 74192 will clear the load. This could be useful in preventing the loading function, if properly planned.

testing

It is possible to feed the input amplifier with 60 Hz from the ac line, an audio

oscillator, a grid-dip oscillator or a tap from the timing chain to test the count chain. A scope together with ac and dc inputs to meters can also be used. The dc assumes a medium value except that it is lower in the D output of decades due to the duty cycle in the divide-by-five section. The scope distinguishes between true stable counting and added noise counts. A low-C probe for the scope and an rf probe for the meter are necessary at high frequencies. However, the scope can often be put at the end of a decade or two to show what is going on at high rf inputs. Above 20 MHz of so, there does not seem to be much of a square wave anywhere.

digital dials

W7TX has covered the method of *synthesizing* the original transmitting or receiving frequency by heterodyne mixers. In this method, the three frequencies in the Collins S-Line are mixed in a manner that results in up-counting the synthesized operating frequency itself.

In general, this method is accurate, although you can expect one small difference: When in the transceive mode, the transmitter is on a slightly different frequency if the bfo is not on the same frequency in the two units – due either to the receiver's variable bfo or to the fact that the two bfo crystals may not be on exactly the same frequency. This would show up if the counter were switched from the receiver to the transmitter.

In some types of reception, the count accuracy depends upon zero beat. This is satisfactory with ssb, but involves some possible error on CW unless some means of limiting the error is employed. In correspondence W7TX pointed out that sufficient bfo signal leaks into the i-f so a scope on the i-f output would indicate when there is zero beat between the signal and the bfo. Also, the "signal mode" by which the i-f signal itself is counted, rather than the bfo, is often preferred by the user.

When the signal is keyed, the counting in the signal mode may be a longer period than the dashes, thus giving many short counts unless the key is held down. This happens particularly in the high resolution (long counting period). On the other hand, while the read-out will often stand still in the lowest resolution (short counting periods) the decimal point then is moved over so that the frequency is not measured as accurately.

If an oscillator, such as the exciter's CW calibrator, is placed in zero beat with the signal, or to one side of frequency-shift keying, accurate information on frequency and shift can be obtained.

Somewhat more design work may be required for using the synthesis method in some receivers[3] such as those where the heterodyning process changes on different bands. In some others, such as the Racal,[4] it may prove to be almost impossible to filter the resultant desired frequency from all its components.

It has been mentioned in the Motorola application note that the MC1496 and μA796 double-balanced mixers can operate as frequency doublers by introducing the same signal at both input ports. Such a device might provide a convenient way to multiply oscillator frequencies to separate several receiver or exciter oscillators.

The CA3001 amplifiers are covered by an RCA data sheet, File No. 122, and an application note, ICAN-5038. Although many of the circuits shown require a negative as well as a positive power supply, the small negative current can frequently be obtained from the same power supply.

The cost of the CA3001 is higher than that of the MC1550. The CA3001 differs from several competing types due to the incorporation of emitter-followers. These permit cascading stages with similar connections. Useful gain can be obtained beyond 30 MHz although the design goal was 20 MHz in a broadband amplifier.

Tests of the low-cost MC1550 provided negative gain when I attempted to cascade two of them. Motorola Application Note AN-299 discusses this, and proposes that the second of a two-stage video amplifier be operated as a com-

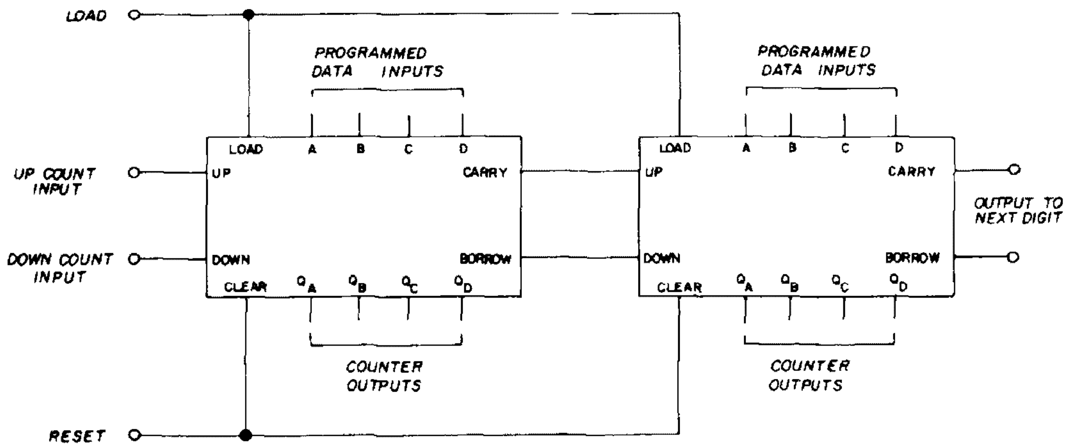


fig. 6. Method for using serial up/down counters, type 74192.

mon-collector, common-base amplifier. This permits direct capacitive coupling between stages.

Both of these devices can provide symmetrical limiting, thus tending to square up the waveform when the low input requirements are exceeded slightly.

Another matter should be kept in mind when recreating the original operating frequency. The recreated original frequency should not create interference to the receiver. Also, the transmitter output should not cause troubles in the counter.

In arranging the switching in the station digital accessory, provision was made for use of the synthesis method, and also for putting a vfo signal through a tuned circuit and separate amplifier before connecting it to the counter input amplifier. In practice, however, the Racal and Collins receivers produced adequate voltage directly at the counter amplifier without the use of a tuned circuit or special amplifier. However, such an amplifier should increase the isolation between the receiver and the counter, which would be desirable.

up/down counting

One goal for the digital station accessory was to provide for a direct vfo count with programmed frequency corrections. In that way, the dial of essentially any receiver or exciter can be replaced with a digital electronic counter. One means to do this is with the 74192 up/down counter.* Like the 7490 decade, it takes a plus voltage to reset. It has two inputs and outputs, one input to be driven while the other is held at a logic high. It has four binary-coded decimal inputs for presetting when a *load* terminal receives a logic low pulse. These BCD inputs, when not connected, will assume a logic high, so they must be grounded if a load pulse is not to transfer the BCD inputs to the internal decade counter. It is shown in **fig. 6**. As stated, the up/down counter actually is a frequency divider like the 7490 and programs always go high, but only the BCD read-out is down-counting.

The digital accessory, as now built, continues to use the A section of a 7490 before the count gate, to obtain suitable performance, followed by the BCD section (see **fig. 3**). This means that the units read-out will actually be up-counted, never down-counted. However, if the units figure is required, all is not lost. It is necessary only to subtract it from zero for the actual frequency. For example, say that a frequency shows as 14,200.794. Subtract the ending 4 from 790, and the actual frequency is 14,200.786. This is subject to the plus or minus one-count error of the count gating and the A section of the first decade.

Also, only five up/down counters are

*Although not advertised at a comparable price, the one-line-input and up/down control is available in the SN74190N. The programmable SN74196N decade is also useful, but it is an up-counter only.

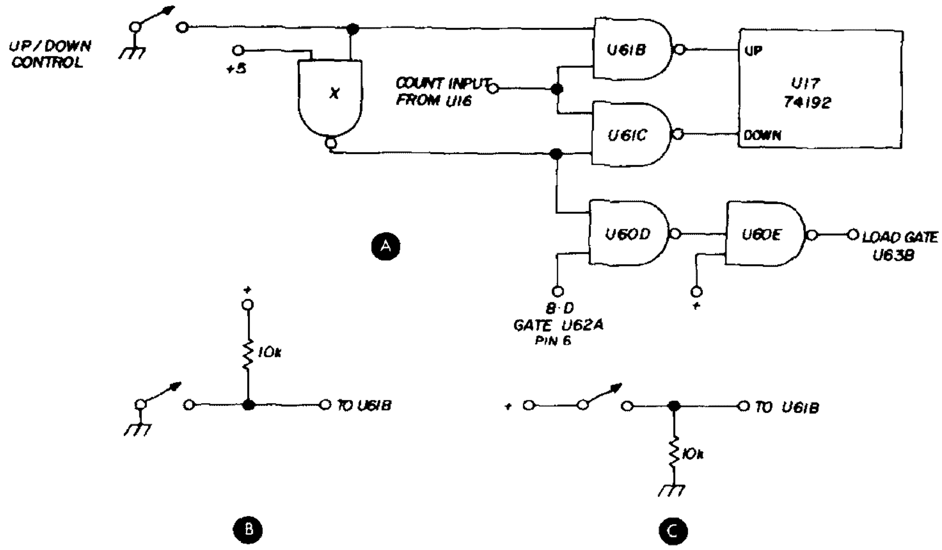


fig. 7. (A) Proposed method of single-line control of up and down count, with automatic elimination of load pulse for up-counting, if desired. Note that U60D can be fed from a switch as a single-line control. (B) Method of forcing gate inputs high if they do not assume a high level. (C) Method of using a switch on positive instead of negative power if desired for switching convenience.

used. One reason is that it is not usually feasible to count the megahertz in vfo counters, although arrangements could be made for a dc-switched display of megahertz. The carry from the sixth read-out to the seventh and eighth, therefore, will be up-counted. It is convenient to cut them off.

Ordinarily, the seventh decade could be driven by the D output from the last up/down counter. However, *if it is driven* from the carry output of the up/down counter, then the seventh and eighth read-outs will operate only on up-counting, but not on down-counting, which is desirable. This fits the specific application for the Racal receiver, and may fit others as well because the vfo itself does not indicate the band involved.

When tapping signals from the receiver oscillator or mixer, take care to avoid noise, birdies or transmitter pickup on the lead. RG-62A/U coax has been recommended for its lower capacitance than RG-58B and RG-174. For testing purposes, a Vector Voltage and Current Test Adapter can be inserted under the desired tube. It permits attaching a resistor in series with the coaxial cable. The resistor frequently can be as large as 5k to 7k. The coax feeds the counter's input amplifier through a selector switch. Should the signal not be sufficient, presumably a CA3001 or MC1550 IC video amplifier, possibly with a tuned circuit, could be inserted.

If the one-second display is not too far behind the tuning dial rotation rate, it is the most satisfactory resolution. Next would be the 10-Hz resolution, producing a one-tenth second count; this takes one less up/down counter, but removes all doubt as to the nearness of the indicated frequency to the 0.1-kHz read-out. In any event, you may make switching provisions to remove V_{cc} or ground from the decoder/drivers and read-outs beyond the effective count digits for the application involved, should they show up and be disturbing to the accurate reading of the dial.

Usually there are several ways to do the switching and control. The up/down counters must have plus V_{cc} on the unused up or down input. One way to do this is to have two gates, one for up-input and one for down-input, so arranged that a ground on the control-line second input of a gate will make that one's output go high (see fig. 3). This requires two control lines and, sometimes, more than minimum switch facilities.

A single control line, either for V_{cc} or ground, probably can be devised. One

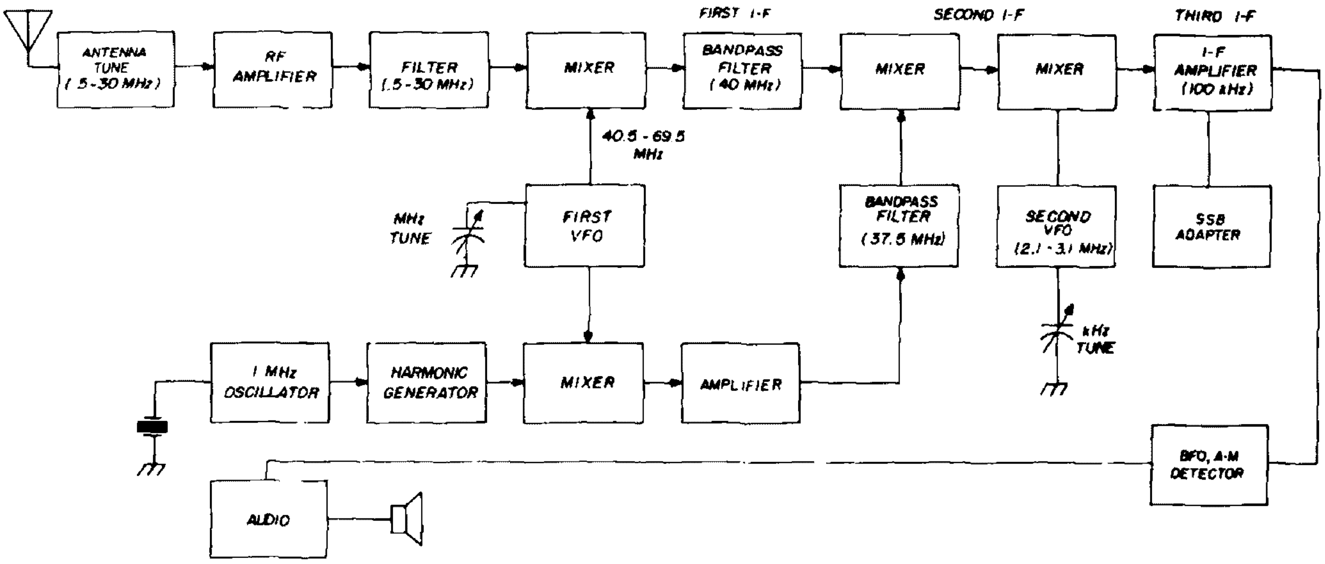


fig. 8. Block diagram of the Racal RA-17C.

way is to place a NAND inverting gate between the second inputs of the two up and down counting gates as shown in fig. 7. If a gate input does not assume an adequate positive potential when left open, a 10k resistor can be provided so that one control line can handle either ground or V_{CC} to switch between up and down counting. One of these may permit using the same switch contacts as already are serving another purpose, such as changing the programmed loading for several bands.

At my station only up-counting is required when not using a programmed error correction, and only down-counting when loading a program. These are suitable for Collins equipment (which can be through a mixing unit) and the Racal (on a direct vfo down-count). The elimination of the loading pulse can be accomplished automatically with a suitable AND gate in series with it, or two NAND gates which will reverse twice.

There are other ways of applying a gate to minimize the number of control lines and switch poles. One way is to use a 4-input load gate at U63B. Another is shown in fig. 7.

It is convenient to use the AND gate or some left-over NAND gates to cut off any one of the coincidence gate inputs to U63B that select the position of the load pulse. One consideration is to feed only a logic low to a following NAND gate to cut it off – a logic high will wash out the input pulses and feed the following gate with a logic high so it is not cut off.

programming

Let's say that we wish to down-count a vfo, and to have the counting start at a programmed frequency which will correct for all other oscillators and errors in the receiver. This is not difficult.

After a long warm-up, set the receiver dial exactly in zero beat at the zero end of the dial, with WWV or a harmonic of the digital station accessory's crystal calibrator after careful adjustment to WWV. Then, in the up-count position, determine the reading. This becomes the amount to preset into the up/down counters. Each digit should be converted to its binary equivalent, and the ABCD inputs of the 74192 ICs grounded except where they are to be at a logic high as required for the binary equivalent for that digit. Do this for each digit.

Example: The Racal receiver covers the entire 30-MHz range, with or without a tuned input circuit, and beats it with an unstabilized oscillator in the 40 to 70 MHz range. This vfo also feeds a mixer which mixes harmonics of a 1-MHz crystal, passes the mixed result through a 37.5-MHz band-pass amplifier and back to the signal stream. There, a mixer

preceeded by a 40-MHz bandpass filter mixes the 37.5-MHz signal to produce a 2- to 3-MHz second i-f (see **fig. 8**).

Since the unstabilized oscillator is injected and removed (at another harmonic of the 1-MHz crystal) the second i-f signals have crystal accuracy. At this point, a 2.1- to 3.1-MHz vfo tunes the signal, and is counted, before it produces a 100-kHz IF. This, in the RA-17C at least, is mixed with an 82-kHz oscillator fitted with a fine-tuning dial, passes through L/C sideband filters in the 18-kHz range and is detected.

The thing to remember here is that the 2.1- to 3.1-MHz vfo covers a 1-MHz band and contains an error of at least the 100-kHz of the following i-f.

Set the dial at zero (vfo output, 3.1 MHz). Using one-second gating, count the vfo frequency. We can discard the megahertz figure because it will be beyond the first six digits and will not appear on the display, but we must program the 100,000-Hz i-f error and any other little errors lying around. This is done quite simply – by grounding every ABCD data input to the up/down counters except the A input to the sixth decade (fifth up/down counter in this case). This means that the six digits after a reset and load will read 100,000 *before* counting.

Now, for one second, this is down-counted (disregarding the 3 MHz which will not show), and the 3.1 MHz signal reduces the 6-digit reading to 000,000 Hz. This is what we want.

At the 2.1 MHz end of the dial the down-counting would fall 1 MHz short and leave it at 1,000,000 Hz. But, again, the 1 MHz is not indicated. All frequencies between the vfo ends are shown to the Hertz.

If the vfo to be counted has an increasing output frequency with its dial settings, then a slightly different procedure is used. Loading the program results in the same upward indication. But, we need a program below zero, for up-counting. Therefore, turn the vfo to the low-frequency end of the dial, set it carefully to zero beat with WWV or a calibrator harmonic, and count the frequency. This count must be subtracted from zero, and the result wired into the program.

Again, let's take an example based on a fictitious reversal of the Racal dial used in the down-count example above. Assume that zero frequency produces a 2.1-MHz vfo output, while the 1000-kHz higher end of the dial produces a 3.1-MHz output. Assume that the count at the low end is actually 2100 kHz. Here again, it would be desirable not to indicate megahertz through counting spill-over into the seventh and eighth decades. So, we have 100,000 to program. We subtract this from zero; the answer is 900,000 (to six figures). In the sixth count decade – in this case the fifth up/down counter – the A and D inputs should be left open or connected to a logic high, and all other inputs should be grounded.

Let's examine the results. The oscillator is at 2.1 MHz, so it starts up-counting at 900,000. The first 100,000 Hz brings it to an even 1 MHz, and the next 2 MHz brings it to a total of 3 MHz. But we decided not to indicate the MHz reading by a counting means, so the read-out is 000,000 Hz. Similarly, at the high-frequency end, the 900,000 programmed figure is raised by 3.1 MHz to a total of 4 MHz, and again it reads 000,000 Hz, inasmuch as we do not indicate the 4 MHz.

The Racal's fine-tuning control on the stable 82-kHz oscillator makes it possible to use the 100,000 program in every case, and then to adjust the zero setting on WWV with the fine tuning, which holds for every band.

To avoid errors due to the Racal's 1-MHz crystal being slightly off frequency, remove the 1-MHz crystal and feed the oscillator with 1-MHz obtained from the counter's 1-MHz test output, plug 5 of the input and time-base board as shown in fig. 2, last month.

Soldering does the job when only one band is to be programmed (30 bands in the Racal). In many receivers, however, five or more bands, involving different errors, may be present.

The bandswitch on the digital station accessory can be set to the same band as

table 1. Plug assignments common to two or more boards.

plug	to	purpose
D	RO-C-IT	reset 7473
E/J	reserved	program
K	BS3e	program
L	all	transfer invert
M	RO-C	count in/out
NPRS	IT-RO	tens BCD
TUVW	IT-RO	units BCD
Y	power	+12 V_{cc}
4/7	RO-C	100k BCD
8/11	RO-C	10k BCD
12/15	RO-C	1k BCD
16/19	RO-C	100-Hz BCD
20	IT-C	load 74192
21	IT-C	carry/up
22	IT-C	borrow/down

the receiver or transceiver. This switch can provide a pole for selection of up or down counting, provision of the load pulse and selection of the correct program for the band. Instead of soldering the programmed inputs in place, all programs should be counted for several bands, and written down. Then, convert these to binary-coded decimal for each digit. Add the values of the A, B, C and D outputs (1, 2, 4 and 8) to convert.

If all bands require a particular input pin to be at a logic high, that pin is left open. If all bands require a particular data input to be at a logic low, the pin must be grounded. But many inputs will have to be high on some bands, and low on others. For those bands on which a specific input must be grounded, this can be done through a steering diode (they can be mounted perpendicular to the *Plugbord* in nearby unused holes) for each affected band, with anode to the pin and cathode to the bandswitch contact. Then, when the bandswitch is turned to a particular band, it will ground all diodes connected to that contact.

For a particular five-band equipment, as many as four diodes (such as surplus silicon small-signal diodes) may be necessary to connect a particular input pin to the switch contacts for several bands. In the equipment described, with five up/down counters, each with four inputs, as many as fifteen diodes could connect to a single band-switch contact. On the average, however, somewhat less than half of this number may be necessary for the required programs. Five board plugs, E through J, have been reserved to ground the program input pins.

If the same bandswitch contacts are to be used for grounding some other circuit, such as up/down control, a steering diode may be required in this additional line to prevent the assumed logic high on the input pins from affecting others.

For ultimate cycle-accuracy, much of the above depends upon determination of zero beat. Sometimes, particularly with high-fidelity earphones or a loudspeaker, this can be done by ear. W3FQJ has described how to use a scope to show zero beat when connected to the i-f.[5] Another way is to use a separate audio oscillator, and adjust WWV and the vfo under test to the same audio beat, and then correct for the indicated offset oscillator frequency. A source of this audio signal is the 1-kHz output from the time-base chain of ICs, preferably through a stopping capacitor or high value resistance. When WWV or any other signal is in phase with this 1-kHz signal on the scope, the vfo is exactly 1 kHz off zero beat, and the read-out can be corrected mentally.

wiring

The time-base wiring was described last month. The remaining wiring for the input amplifier, resolution, gating, control and count decades can follow the accompanying figures and tables. These are based upon separate up/down controls, and taking the load-pulse control line out to a switch. Methods of reducing the plug and switch requirements for these have already been discussed. Addi-

table 2. Additional plug assignments for the input and time-base board.

plug	to	purpose
12	RS4b	1 kHz
13	RS4c	100 Hz
14	RS4d	10 Hz
15	RS4f	resolution
18	BS2c/j	up count
19	BS2b	down count
X	MS4a	rf in
M	BS3a	load control

table 3. Additional plug assignments for the read-out board.

plug	to	purpose
X	RS2b	decimal U32
21	RS2c	decimal U33
22	RS2d	decimal U34

tional plug assignments are given in tables 1, 2 and 3.

Connect all V_{cc} pins to the V_{cc} supply. Bring a NAND gate input, if unused, to V_{cc}. Bring all ground pins to the ground bus. In the final divider and the gating, follow fig. 5. Proceed with the detailed board connections listed in table 4 for the input and time-base board. Note that +5-V buss for both plug-A and plug-B, on each side of the broken foil, must be connected to their proper plugs.

The count-board connections are given in table 5. Make similar connections from decade dividers to memory latches in each decade. The latch outputs can be connected more easily to the AND gates from the timer FFs than directly to the indicated plugs. However, there are three places where there is no time output from the time FFs to the plugs due to minutes and hours being less than 100. With counters being fed V_{cc} through plug-A only when counting, and the AND gates receiving +5V from plug-W only when time is displayed, the outputs can operate in parallel as indicated.

The read-out board connections appear in table 6. Connections for eliminating leading-edge zeros have been discussed; other connections involving the decoder/drivers and the read-out units are the same for each digit. Note, however, that the two unused inputs to the final memory latch should be grounded (or the same inputs to the decoder/drivers should be grounded) to prevent arbitrary activation of the final read-out above the digit 3. It will be recalled that plug-B is used for +5 volts to the middle four digits for use as a digital clock, and the two digits at each end receive their +5 volts from plug-A.

Switch wiring, using a new 4-pole, 11-position bandswitch for controlling band, up/down count, loading and load pulse is tabulated in table 7. No specific assignments are given as yet to the plugs for the mixer board, which has not yet been wired and tested with the shielded Weeductors and the MC1550 amplifier. Should anything of unusual interest develop in its construction, supplementary material will be published in *ham radio*.

To minimize the chance for switching that might step up the digital clock, a filter capacitor was added from the +5-volt switchable power supply to ground. Also, several 10-μF capacitors (or larger) were put across the resolution switch contacts to ground whenever they in-

table 4. Connections on the input and time-base board.

from	to	purpose
Plug-X	amp input	rf in
Plug-A	V_{cc} bus	+ count
U16p14	amp out	count in
U16p2	V_{cc}	unused
U16p3	U62Cp8	reset
U16p7	ground	R9
U16p12	U61Ap4	A out
U16p1	U61Ap6	BCD in
U16p9	U18p3	B out
U16p8	U18p6	C out
U16p11	U18p7	D out
U18p16	Plug-T	A out
U18p15	Plug-U	B out
U18p10	Plug-V	C out
U18p9	Plug-W	D out
U18p2	ground	A in
U61Bp1	Plug-18	up
U61Bp3	U17p5	up
U61Bp2	U61p11	D out
U61Bp2	U61Cp12	D out
U61Cp13	Plug-19	down
U61Cp11	U17p4	down
U17p14	U16p3	reset
U17p11	U63Bp12	load
U17p13	Plug-22	borrow
U17p12	Plug-21	carry
U17p3	U19p2	A out
U17p2	U19p3	B out
U17p7	U19p6	C out
U17p6	U19p7	D out
U17p7	Plug-N	A out
U19p16	Plug-P	B out
U19p10	Plug-R	C out
U19p9	Plug-S	D out
U62Dp12	V_{cc}	transfer
U62Dp13	U63Cp8	transfer
U62Dp11	U19p4/13	transfer
U19p4	U18p4/13	transfer
Plug-Y	+12 bus	amplifier V_{cc}

table 5. Connections on the count board. See text regarding similar connection to each decade.

from	to	purpose
Plug-A	V_{cc} bus	+ count
Plug-21	U22p5	up
Plug-22	U22-4	down
Plug-D	U70Cp13	reset
U70Cp11	U22p14	reset
U22p14	U23p14	reset
U23p14	U24p14	reset
U24p14	U25p14	reset
U70Cp12	V_{cc}	unused
Plug-L	U70Bp4	transfer
U70Bp4	U70Dp10	transfer
U70Bp5	V_{cc}	unused
U70Dp6	U22p4/13	transfer
U22p4	U23p4/13	transfer
U70Dp9	V_{cc}	unused
U70Dp8	U24p4/13	transfer
U24p4	U25p4/13	transfer
Plug-20	U22p11	load
U22p11	U23p11	load
U23p11	U24p11	load
U24p11	U25p11	load
U22p13	U23p4	borrow
U22p12	U23p5	carry
U23p13	U24p4	borrow
U23p12	U24p5	carry
U24p13	U25p4	borrow
U24p12	U25p5	carry
U25p12	Plug-M	count out
U22p3	U26p2	A out
U22p2	U26p3	B out
U22p6	U26p6	C out
U22p7	U26p7	D out
U26p16	Plug-16	A out
U26p15	Plug-17	B out
U26p10	Plug-18	C out
U26p9	Plug-19	D out
U27p9	Plug-15	D out
U29p10	Plug-6	C out
U29p9	Plug-7	D out

volved breaking the +5-volt supply to the AND gates on the count board, breaking the supply to the time decoders and read-outs and breaking the +5-volt supply for counting.

potpourri

The possibility of rf interference to the counter has been mentioned. A completely closed cabinet or chassis might be useful, but may not be effective unless the 117-volt ac supply and other leads from the unit were suitably treated. In the presence of substantial capacity on input and output circuits, and IC count input and output circuits do not always toggle properly.

Interference from the counter was reduced materially by placing a 0.02-μF capacitor directly across the input of each LM309K and LM336 voltage regulator. Interference was also reduced by reversing the power plug – which suggests that a statically shielded power transformer might be useful. There was also some leakage of 1-MHz harmonic output, modulated nearby by 1 kHz, which did not occur in the earlier counter which used a double switch to prevent leak of the calibrator signal to the receiving antenna.

Shielding is needed to avoid addition of receiver birdies and other noise into a counted oscillator. This is more likely to occur with the direct up/down counting method (unless an IC amplifier reduces it)

table 6. Connections on the read-out board.

lug	to	purpose
V_{cc}	Plug-A	+5 volts
ground	Plug-Z	-5 volts
Plug-M	U52p14	7490 in
U52p12	U52p1	A-BCD
U52p11	U51Ap1	count
U51Ap12	U51Bp5	count
Plug-D	U80Ap1	reset
U80Ap1	U51Ap2	reset
U51Ap2	U51Bp6	reset
U80Ap2	V_{cc}	unused
U80Ap3	U52p2	reset
Plug-L	U80Bp4	transfer
U80Bp6	U53p4	transfer
U53p4	U53p13	transfer
U53p13	U54p4	transfer
U54p4	U54p13	transfer
U52p12	U53p2	A out
U52p9	U53p3	B out
U52p8	U53p6	C out
U52p11	U53p7	D out
U51Ap12	U54p2	A out
U51Bp9	U54p3	B out
U54p6	ground	no C
U54p7	ground	no D
U53p16	U45p7	A in
U53p15	U45p1	B in
U53p10	U45p2	C in
U53p9	U45p6	D in
U51Ap12	U46p2	A in
U51Bp9	U46p3	B in
Plug-K	U35p9	decimal
Plug-22	U34p9	decimal
Plug-21	U33p9	decimal
Plug-X	U32p9	decimal

table 7. Wiring connections for the mode switch and bandswitch, including provision for synthesized mixing board and five-band programmed correction.

lug	to	purpose
MS1a	M plug	hf mixer
MS1b	phono	transmitter hf
MS1c	phono	receiver hf
MS1d	MS1c	receiver hf
MS2a	M plug	vf mixer
MS2b	phono	transmitter vf
MS2c	phono	receiver vf
MS2d	MS2c	receiver vf
MS3a	M plug	bf/if mixer
MS3b	phono	transmitter bf
MS3c	phono	receiver bf
MS3d	phono	receiver i-f
MS4a	IT plug-X	rf in
MS4b	M plug	amplifier out
MS4c	MS4b	amplifier out
MS4d	MS4c	amplifier out
MS4e	phono	Racal vfo
MS4f	phono	counter
BS1a	M plug	mixer out
BS1b	M plug	2 - - 3 MHz
BS1c	M plug	3.5 MHz
BS1d	M plug	5 MHz
BS1e	M plug	7 MHz
BS1f	M plug	10 MHz
BS1g	M plug	14 MHz
BS1h	M plug	21 MHz
BS1i	M plug	28 MHz
BS2a	ground	control
BS2b	IT plug-19	down count
BS2c/j	IT plug-18	up count
BS3a	IT plug-M	load control
BS3b	open	load
BS3c/j	ground	no load
BS4a	IT plug-K	program
BS4b/J	reserve	load

than with the synthesis of the operating frequency by the mixer method.

Since building the unit and writing this article, I have found that the power input to the LM309K regulator that most required the 0.02-μF capacitor across the input had a substantial saw-tooth wave. This mostly was eliminated in the regulator but there were still some small spikes in the output. These may have been caused by the input waveform, or may have come down from the counter, though synchronous with 60-Hz power.

In any event, while the spikes did no harm when counting rf, they occasionally caused irregular counts at audio frequencies. In fact, I suspect that this noise is responsible for the problem described above in gating the very first FF of the counter, which led to the unusual gating circuitry. Also, it appears to place a limit on how much coaxial cable capacitance can be present on the interval-timer board, to the switch, to the phono plug and beyond the chassis to the receiver. While RG-62A/U cable reduces capacitance, so does the length of the cable. This will be investigated further, with a view to putting the count gate ahead of the A section of the first count decade.

If the bypass capacitors do not eliminate the tendency of switching spikes occasionally to step up the digital clock, to consideration can be given to using a shorting-type switch. If the AND gates cause a spike in the V_{cc} line when it is switched on and off the AND gates, then these gates can be controlled with an input line while leaving the V_{cc} on them.

conclusion

There have been many tests, experiments, substitutions, and the like, in the construction of this unit. However, the results have been worthwhile, making a pleasant addition to operating the station. The unit has been a help in Official Observer and Intruder Watch work, particularly in the ease with which accurate frequencies and fsk shifts can be determined. As stated initially, the facility with which digital IC equipment can be built makes them particularly adaptable for homebrew and self-designed items in the amateur station.

references

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ham radio

audio-actuated in-line squelch... the squelcher

A new twist
to an old idea
that can be added
to any receiver
in one evening
without
major modification

Gene F. Greneker, K4MDG, 2546 Wilson Woods Drive, Decatur, Georgia 30033

While the avc-operated squelch has been around a long time, there's something to be said for the audio-actuated in-line squelch, and the Squelcher is the last word in in-line squelch circuits! Traditionally, the squelch circuit has been used to silence receivers during inactive periods. The Squelcher has other important functions other than receiver muting during periods of inactivity. For example, the Squelcher makes possible clear channel operation even in 20 dB of noise, if the received signal is at least 5 dB above the noise level. There will be some noise under the received signal, but the ear, an amplitude-type detector, hears the audio output of the receiver go from a zero level during the squelched period to a maximum during transmission; the effect is to create an auditory illusion. The desired received signal seems much stronger than it actually is. In short, the Squelcher can mask a great deal of background noise between words and sentences that contributes to operator fatigue during long operating periods.

The Squelcher is basically an audio-actuated switch that closes a set of relay contacts in series with the speaker each time the receiver output exceeds a pre-set level. As long as the audio level stays above the pre-set threshold, the relay stays closed, keeping the speaker connected to the receiver output. However, when the receiver output falls below the pre-set level, and after a time delay determined by the time constant of a capacitor (C3), the relay contacts open and silence the speaker.

While most squelch circuits rely on a change in avc voltage to achieve squelch action, the Squelcher operates on the change in the audio output of the receiver. The addition of the agc squelch circuit to an existing receiver can be a custom project with as many variations in circuitry as there are different makes of receivers. The Squelcher, on the other hand, can be added to any existing

receiver with an output impedance of 4 or 8 ohms and will achieve the same results.

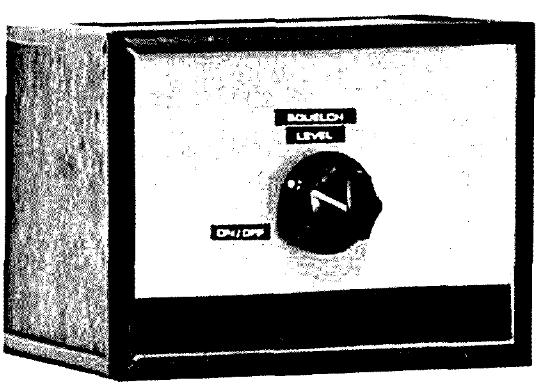
theory

The Squelcher consists of an internal rectifier bridge that converts 6.3 Vac to dc to operate an amplifier stage and a relay driver stage. Diodes CR1, CR2, CR3 and CR4 comprise a standard full-wave bridge rectifier whose output is filtered by capacitor C4. Resistor R6 serves to limit the current flowing through zener diode (CR7) to a bias point that allows the zener to regulate the voltage to 8.2 volts regardless of load.

Transformer T1's primary is connected directly across the receiver output. During periods when the threshold is too low to allow the receiver output to be switched to the speaker, T1 presents a constant 8-ohm load to the receiver. When the output rises to a level sufficient to switch in the speaker, the low-impedance input of T1 keeps the reference level to the amplifier fairly constant regardless of the additional speaker load.

The secondary of T1 is fed into the high side of the potentiometer (R7) which serves as the squelch or threshold-level control. Capacitor C1 passes the ac component into the 2N2925 amplifier stage, which amplifies both positive and negative peaks in linear fashion. The output of the transistor amplifier stage (Q1) is coupled through C2 to a negative peak clipper (CR1) and a half wave rectifier (CR2) whose output is averaged by capacitor C3. The value of C3 determines the time constant or hold-in time of the relay driver circuit. For shorter hold times, decrease the value of C3. The voltage stored by C3 is applied to the base of transistor Q2 through resistor R5. When the voltage applied to the base of Q2 reaches a sufficient level to cause conduction, relay K1 closes, connecting the speaker to the receiver.

Switch S1 is mounted on the threshold potentiometer. In the off position, S1 shorts the open contacts of K1 connecting the speaker directly to the output of the receiver, taking the squelch circuit completely out of the circuit. This action also opens the 6.3 Vac input line, disconnecting the Squelcher from the power supply.



Front view of the Squelcher.

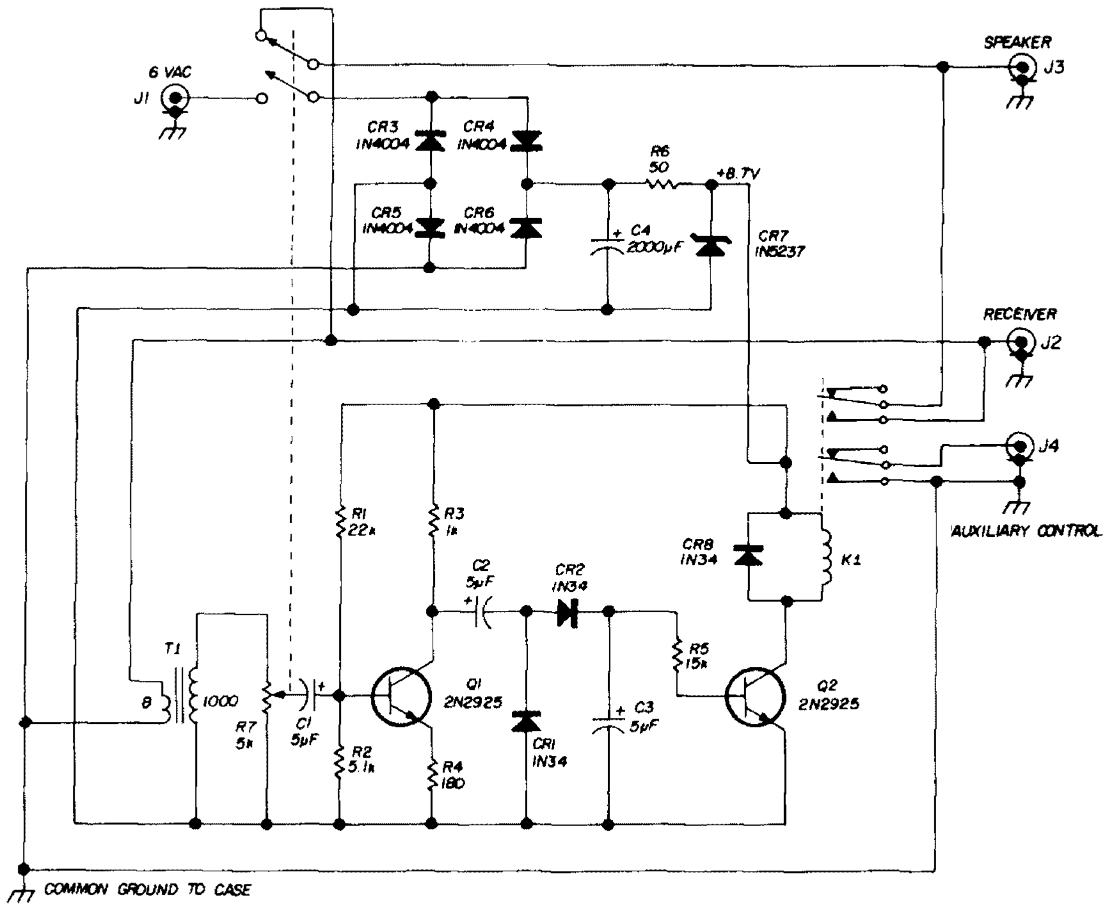
construction

The parts layout is not critical. However, for ease in assembly, it may be best to make your own printed-circuit board or to purchase the ready-made board.* If you use the pc board, parts should be mounted as shown in the photograph, with the exception of capacitor C4 and potentiometer 1 – these will be mounted later. When all the other parts have been mounted, check to insure all leads were clipped close to the foil on the pc board.

Mount the potentiometer first. The three leads of the potentiometer are designed to be soldered directly to the board. Solder the three leads of the potentiometer to the foil side of the board and trim any excess leads. Make sure the potentiometer is *flush against the component side of the board*, or the board may crack during final tightening of the mounting nuts. Next, solder 1½-inch pieces of wire to the switch portion of the potentiometer.

When installing capacitor C4, be care-

*All components, including cut and drilled printed-circuit board, pre-drilled Eico Flexi-Cab cabinet, and the input transformer are available for $19.95 from H & L Electronics, Box 9707, Atlanta, Georgia 30319. The printed-circuit board alone is $2.95, and the transformer is $1.00 from H & L.



K1 dpdt reed relay, 12 Vdc coil (Allied Control RF-2A)

fig. 1. Schematic diagram of the Squelcher. All capacitors are rated at 25 Vdc, all resistors are half watt.

R7 5k, ½ W potentiometer with S1 attached (Clarostat A42-5000K-6140 with type 21 switch attached)

S1 two, spst switches, one on and one off per throw (Clarostat type 21 switch attached to R7)

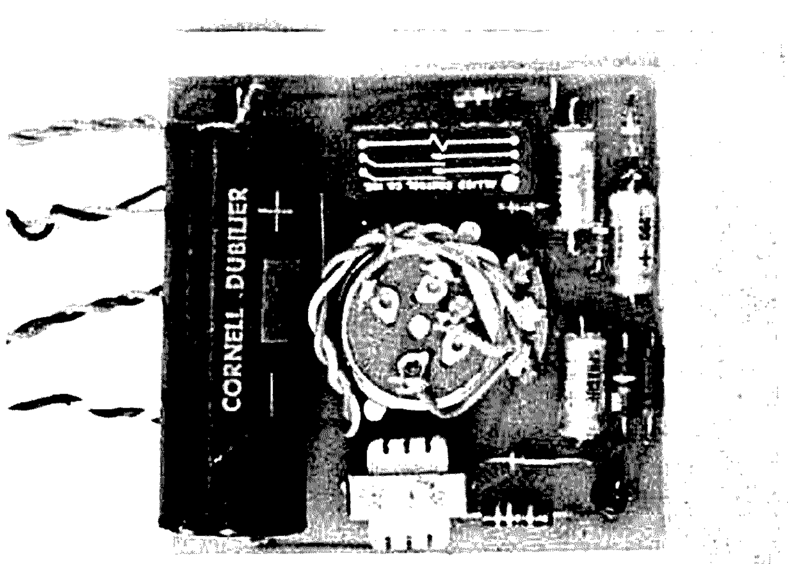
ful to observe the correct polarity. The positive lead goes to the relay side of the board.

When mounting the potentiometer and pc board inside a cabinet, use two lock nuts as spacers to prevent the cabinet from shorting the foil side of the printed circuit. I housed my Squelcher in a 4¼ x 3¼-inch box with the potentiometer mounted in a 7/16-inch hole in the center of the front panel. Construction is completed with the wiring from the pc board to four phono jacks mounted on the back panel. These jacks provide for all interconnections between the Squelcher and the rest of your station.

installation

Two pairs of wires must be brought out of the receiver. Find a pilot light or other source of 6.3 Vac and attach the leads to a phono plug to mate with J1. Be certain to connect the ground side of the plug to a good receiver ground. If your receiver filaments are 12 volts, change R6 to 330 ohms.

Next, the output leads that normally go to the speaker should be connected to a phono plug to mate with J2. The station speaker should be connected to J3 through another mating phono plug. I recommend using shielded cable for these connections to insure a good common ground throughout and to minimize hum pickup. Through J4, you can connect any other unit which you want to control with the normally open contacts of the relay.



The pc board wiring. The potentiometer must be mounted flush against the printed-circuit board.

With the Squelcher connected to your receiver and speaker turn the receiver on. Assuming there is no problem, turn the volume up to a point just above the level that is usually considered comfortable.

Next, turn the Squelcher threshold control clockwise just past the point where the switch clicks on. Tune to a quiet frequency and advance the threshold control until the speaker becomes silent. At this point the receiver is squelched. Tune across the band until a signal shows on the S meter, simultaneously the audio should be heard in the speaker. By varying the setting of the threshold control, the minimum level required to break the squelch can be controlled quite accurately.

additional hints

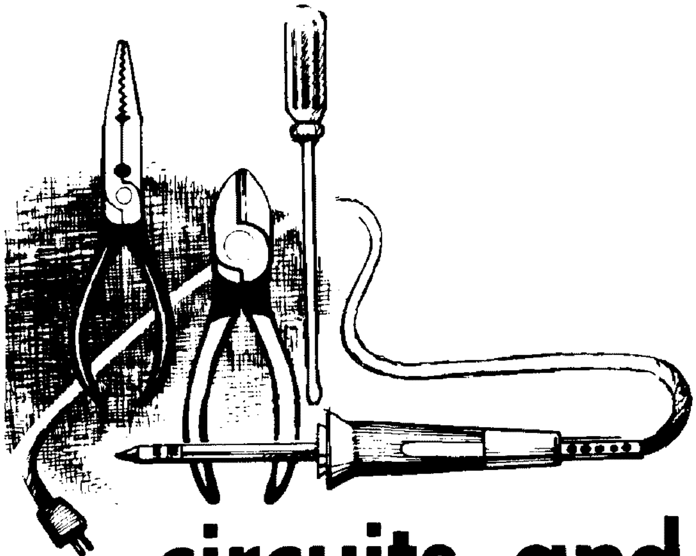
Receivers having agc circuits with a dynamic range of 30 or 40 dB will cause some problems with the Squelcher unless the rf gain control is turned down to a point where the agc action is not noticeable. Otherwise, a signal reading S-9 on the S meter will have the same relative audio output level as a 40 dB over 9 signal. Since the Squelcher detects audio rather than signal level, the audio must be relative to signal strength. Therefore, by setting the receiver audio output for a comfortable level and retarding the rf gain control, the Squelcher will operate dependably.

The optional contacts of the relay will find numerous uses. By changing capacitor C3 to a lower value the reed relay can be driven by moderate to fast CW. By connecting the spare contacts of K1 to a code practice oscillator it is possible to obtain QRM-free copy on strong CW signals, by keying the oscillator in repeater fashion. No doubt, other uses have already come to mind.

After some playing around with the Squelcher, it will be noticed that extremely fine levels in signal change can be detected. As a bonus, operator fatigue will decrease due to squelched noise and interference during lulls in the contact.

Because the Squelcher is somewhat audio-level dependent, it will take a few minutes to determine the proper balance between the receiver audio and rf gain controls. Once these are set, the squelch level will be determined by the threshold control. With time, the Squelcher will become an interference rejection tool for your receiver equal to the Q multiplier, crystal filter and noise limiter.

ham radio



circuits and techniques

ed noll, W3FQJ

digital integrated circuits

One of the exciting side effects of computor technology has been the involvement of a different mathematical form in electronic circuit design and application. The versatility of the true-false (1,0) concept is phenomenal.

Last month we considered inverter, OR, AND, NAND and NOR logic functions and circuits. The basic Boolean expressions for these functions are:

OR	$A+B$
AND	$A \cdot B$
NAND	$\overline{A \cdot B}$
NOR	$\overline{A+B}$

There are quite a number of Boolean theorems and relations. One of the most important is known as the DeMorgan theorem. In practice this theorem proves the validity of a very common form of digital integrated circuit known as the NAND gate. This one type of integrated circuit includes NAND, AND and OR functions. Furthermore, some typical negative-logic NAND gates can also be operated as positive-logic NOR gates. Thus a single digital IC can provide a number of logic functions.

Recall the NAND circuit symbol and truth chart, fig. 1. When both A and B are true, the output is false. However the signal is logically inverted as indicated by the line above the expression in the NAND function equation:

$$X = \overline{A \cdot B}$$

The NAND function can also be written:

$$X = \overline{A} + \overline{B}$$

It is DeMorgan's theorem that equates these two relations.

$$X = \overline{A \cdot B} = \overline{A + B}$$

There are several ways of proving DeMorgan's theorem. Perhaps the most

A	B	X
0	0	1
0	1	1
1	0	1
1	1	0

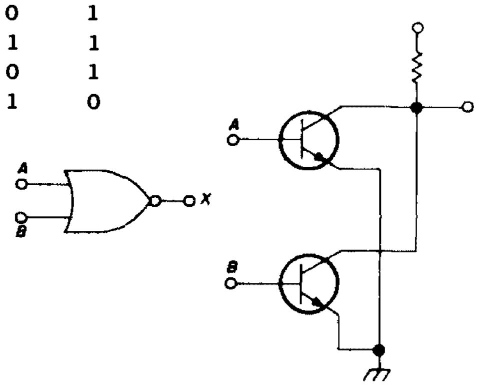


fig. 1. Basic NAND function using negative logic.

obvious uses the truth tables in fig. 1. First we set down the truth table of an OR function. However, the relationship we are trying to prove indicates that the A and B terms have been logically inverted as per the second truth chart. Next set down the truth chart for the NAND function. Note that the output values are identical and therefore:

$$\overline{A \cdot B} = \overline{A} + \overline{B}$$

You can go a step further and invert the logic of the output of the NAND circuit, fig. 2. The fourth chart shows this inversion. Note that the output expression now contains double lines above A and B. These indicate that the signals have been inverted twice. This restores the original logic just as any signal twice inverted is returned to the original polarity. Most importnat, note that the output of the inversion corresponds to the output of an AND circuit. It is readily understandable how a single digital IC can be constructed to include a variety of logic possiblities.

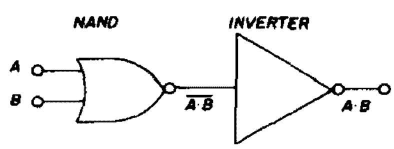


fig. 2. NAND to AND inversion.

A NOR gate, like a NAND gate, can also provide the AND and OR functions. The addition of an inverter (NOT circuit), permits an OR output. The first truth chart of table 2 is that of an AND fuction. An inversion of its logic produces the logic of chart 2. Note that the output is the same as that of a NOR function circuit, chart 3. This proves the validity of the DeMorgan theorem and the following equality can be stated:

$$\overline{A} + \overline{B} = \overline{A} \cdot \overline{B}$$

If the output of a NOR circuit is followed *with an inverter,* you obtain the results shown in the fourth chart. Note that this corresponds to the output of an OR truth chart.

table 1. Validity of DeMorgan's theorem and inversion of NAND to AND function.

OR

$\overline{A}$	$\overline{B}$	$\overline{A}+\overline{B}$
0	0	0
0	1	1
1	0	1
1	1	1

(1)

OR INVERTED

$\overline{A}$	$\overline{B}$	$\overline{A}+\overline{B}$
1	1	1
1	0	1
0	1	1
0	0	0

(2)

NAND

A	B	$\overline{A \cdot B}$
0	0	1
0	1	1
1	0	1
1	1	0

$\overline{A \cdot B} = \overline{A}+\overline{B}$

(3)

NAND INVERTED

$\overline{A \cdot B}$	$\overline{\overline{A \cdot B}}$
1	0
1	0
inverted	AND out
1	0
0	1

$\overline{\overline{A \cdot B}} = A \cdot B$

(4)

In fact, suitable circuits based on the truths of the various Boolean theorems can permit a variety of logic functions. You can use the circuit of fig. 3 as an example. Basically, it is a NAND configuration. Base biasing (resistors R1, R2 and R3) is such that the transistor conducts when logic 1 (positive logic) voltage is applied to both inputs. Under this condition the output will be logic 0. When logic 0 voltage is applied to both inputs or to either input, the transistor is turned off, producing logic 1 voltage at the output. Tied in with a follow-up

table 2. Validity of DeMorgan's theorem and inversion to OR function.

AND

A	B	$\overline{A \cdot B}$
0	0	0
0	1	0
1	0	0
1	1	1

(1)

AND INVERTED

$\overline{A}$	$\overline{B}$	$\overline{A} \cdot \overline{B}$
1	1	1
1	0	0
0	1	0
0	0	0

(2)

NOR

A	B	$\overline{A+B}$	
0	0	1	NOR
0	1	0	in
1	0	0	
1	1	0	

$\overline{A+B} = \overline{A} \cdot \overline{B}$

(3)

NOR INVERTED

$\overline{A+B}$	$\overline{\overline{A+B}}$	
1	0	OR
0	1	out
inverted		
0	1	
0	1	

$\overline{\overline{A+B}} = A+B$

(4)

inverter (NOT circuit), the arrangement can be made to operate as an AND circuit.

The same circuit functions as a NOR gate by changing the polarization of the

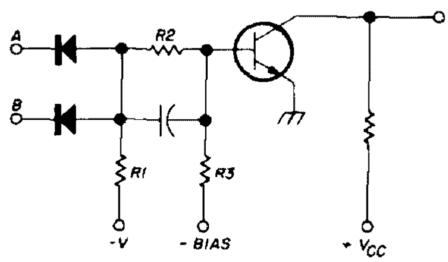


fig. 3. Basic NAND gate.

input diodes and changing the bias level established by the base voltage-divider resistors. In this case, when either or both inputs are at logic 1 voltage, the transistor is turned on to produce a 0 logic output. With both inputs at logic 0, the output is logic 1 because the transistor remains cut off. A follow-up inverter is used to establish an OR function operation.

standard integrated circuits

Digital integrated circuits are found in a number of standardized configurations. The transistor-transistor logic combinations, abbreviated TTL or T^2L, are very common. A basic circuit, **fig. 4**, is a pair of direct-coupled transistors. The input transistor has a dual-emitter, a separate one for each input gate. This circuit operates as a NAND gate.

When logic 0 voltage is present on either or both input gates, the corresponding emitter-base diodes are forward biased. However, no significant collector current exists because of the reverse biasing of the base-emitter of transistor Q2. Stated another way, the saturation current of transistor Q1 is not high enough to forward bias transistor Q2. Therefore transistor Q2 output is logic 1 (positive logic).

The emitter-base diodes of Q1 are shut off when logic 1 voltage is applied to both. In so doing, the collector junction is forward biased. Base current magnitude and direction in Q1 is now such that the base of transistor Q2 is forward biased. Transistor Q2 is now turned on and the collector voltage drops to the 0 logic level.

A complete diagram of an integrated circuit NAND gate is shown in **fig. 5**. It is referred to as a dual 4-input positive NAND gate. Note that two circuits similar to that of **fig. 4** are included in the chip. There are four NOR input gates instead of two. A more elaborate output circuit is *included to* provide higher output to low-impedance loads and to obtain the capability of driving up to 30 loads (high fanout).

I mentioned previously that the AND function can be obtained by the addition of an inverter after the NAND gate. It should also be mentioned that a NOR function can be obtained by cascading two NAND gates. By including multiple gates on an IC chip, any number of common and special logic functions can be established by appropriate external wiring.

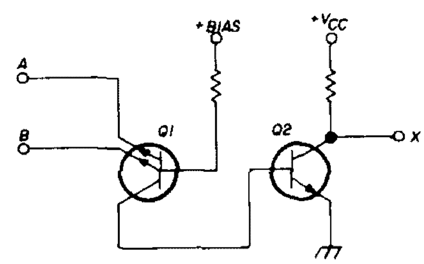


fig. 4. Basic TTL NAND gate.

The diode-transistor logic (DTL) is also common and similar to the TTL ICs. The DTLs are not as adaptable to complex functions and do not have the speed and capability of driving low-impedance loads as the TTLs. The input NOR switches are diodes, **fig. 6**. They are followed by an inverter stage. When both inputs are at logic 1 positive voltage, the diodes are reverse biased. The series or offset diode, however, is forward biased. Current holds the transistor in saturation and its low collector voltage corresponds to logic 0.

The application of logic 0 to any one

or both of the input diodes results in conduction, dropping the voltage at the anode of the series diode to a level that reduces the base current of the transistor to cutoff. Therefore, positive logic 1 output voltage is present at the collector.

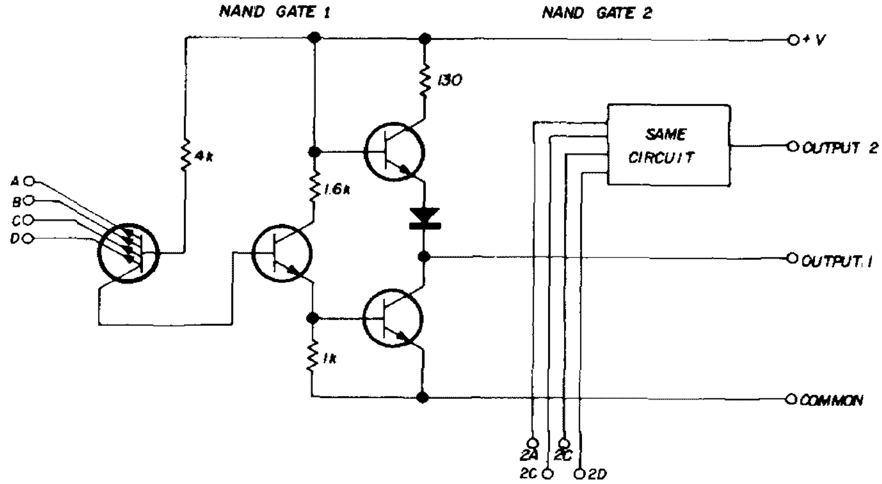


fig. 5. Dual 4-input positive NAND gate (Texas Instruments SN7420).

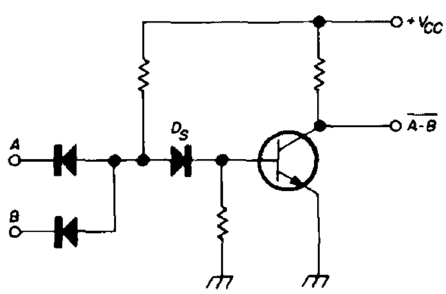


fig. 6. Basic DTL NAND circuit.

A similar family with higher supply voltages and greater power dissipation are known as the high-threshold logic (HTL) group.

Some of the very first IC types were resistor-transistor logic (RTL), fig. 7. This is a basic NOR circuit. These circuits are limited in switching speed and fanout capability. Though economical and adapatle to numerous applications, they have relatively poor noise immunity. Capacitors across the input resistors convert circuits to higher-speed resistor-capacitor-transistor logic (RCTL) types.

An increasingly popular family is the emitter-coupled logic (ECL) digital ICs. These provide the highest speed of all and include such favorable characteristics as low output impedance, high fanout and acceptable noise immunity. The high speed of operation is a result of limited voltage swing in a manner of operation similar to linear types and the non-saturated operation of transistors. These types perform up into the hundreds of megahertz.

In a typical NOR circuit, fig. 8, the emitters connect to a common emitter resistor. This connection prevents saturated operation. Transistor Q4 is used to set the threshold voltage at the input. A succeeding follow-up emitter follower insures low output impedance and good fanout.

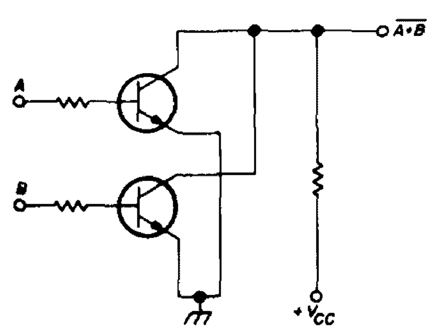


fig. 7. Basic RTL NOR circuit.

The use of metal-oxide semiconductor field-effect transistors (mosfets) in digital integrated circuits adapts them to elaborate, multi-stage and highly-repetitive operations. Complex logic circuits can be incorporated into tiny chips. In a basic NAND circuit, fig. 9, negative voltage establishes logic 1 at the input, and both input gates conduct. The output voltage switches to the logic 0 value. When either or both of the input gates are set to 0 logic voltage, there is an open in the source circuit of the output mosfet and the output is at logic 1 level.

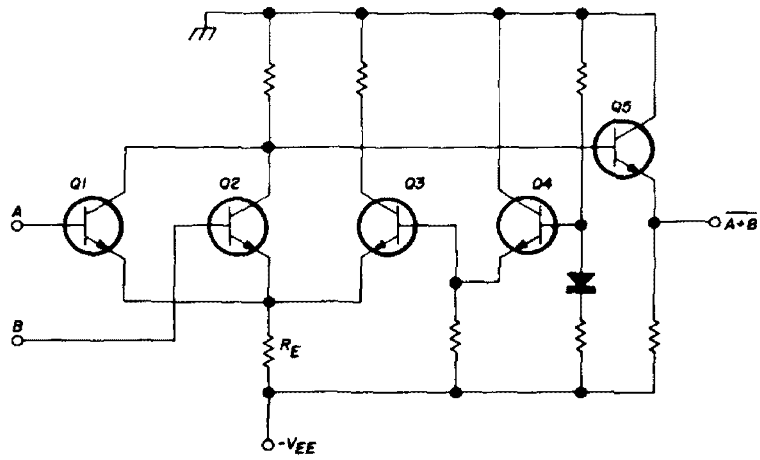


fig. 8. Basic ECL NOR gate.

schedule

The digital IC series will be interrupted next month but will resume the following month with discussions of flip-flops and frequency counters. Instructions will be given on how to set up a small experiment board so you can watch them operate on an oscilloscope and hear them on your receiver. Next month's column will be devoted to antennas to complement *ham radio's* popular antenna issue.

available vfo signal

Here is part of a note from James W. Harrison, Jr., WB4TEX: " . . . I use a Heath SB300 and SB400 cabled up for transceive. Under this arrangement, the receiver feeds its 5- to 5.5-MHz vfo output into the transmitter for frequency selecting, both for receive and transmit. No doubt many other hams use similar setups.

It occurred to me that the receiver vfo should be capable of serving as a tunable oscillator for a QRP transmitter, if such a transmitter included a proper mixer. It appears to me that such an arrangement would make a dandy QRP transceiver setup. I should imagine there are numerous integrated-circuit units capable of handling these functions as part of a QRP CW transmitter . . . "

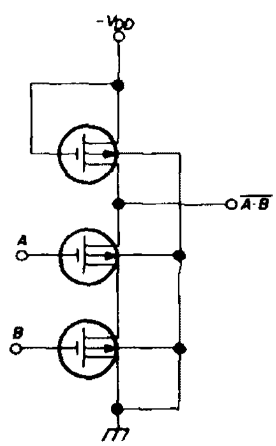


fig. 9. Basic mosfet NAND gate.

Here is a challenge for some of you QRP buffs. Is this not a good opportunity to come up with a vfo-controlled multiband QRP transmitter that you can operate transceive with your big receiver? Thank you, Jim.

ham radio

beam antenna headings

An inexpensive guide to accurate antenna beaming for within the United States and around the world

Irvin M. Hoff, W6FFC, 12130 Foothill Lane, Los Altos Hills, California 94022

Have you ever wondered what direction your beam *really* should be turned to work that elusive station? Have you ever been in a 3-way contact and wondered what heading would catch the other stations on some decent compromise heading? Information is available for beam headings from the central part of the U. S. A. (usually Chicago), and at times from other major cities. This may not help you to any reasonable extent, and the information is pretty much worthless for knowing where to set the antenna for stations here in the U. S. A.

I have wanted accurate beam heading information as much as anybody else, and went so far as to laboriously plot a number of representative cities throughout the U. S. A. on an aeronautical chart. This was quite helpful, and I used the list frequently. I am only a few miles from San Francisco, so the list of foreign beam headings published in the *Foreign Callbook* for that city worked well for me.

While developing a computer-controlled program for aeronautical inertial navigation systems, I decided to use that information to provide a supplementary beam heading program. This is now available, and comes in two lists – the first has 90 domestic cities at well-known places in the United States, the second list has 309 entries of various foreign countries and 10 U. S. cities representing each call district. Thus the program would be useful to stations outside the United States.

The program lists the true heading, magnetic heading, distance in kilometers and distance in statute miles. It thus helps give some indication if the long path is really that much further or not. It is useful to know just how far the other station really is, as it helps give an additional feeling of knowing a little more about the person you are talking with. To further help locate the station, the continent to which it is assigned is listed. If the country is fairly large or has

a best-known city, that is listed also.

Even at a normal computer terminal speed, it takes one hour to print the entire listing of domestic cities and foreign countries. As most computer time costs $20 per hour, there is no way the typical amateur could normally afford to pay for such a customized and accurate print-out. However, I made arrangements to use a high-speed printer, and I have written a fully automatic program. This service is offered to anybody interested for $2 per copy plus 25c postage and handling. Since this would give exact headings from your exact location, this should be of unusual benefit to those interested at a price small enough any amateur could afford. Certainly no profit will be made on this, but so much time has already gone into writing the program and inserting all the data needed, it seemed a shame not to make this information generally available.

I can insert latitude and longitude just as accurately as you can supply it. I already have information on some 30,000 countries, states, cities and towns in the world, but I can just as easily insert the exact coordinates of your very home, if you take the time to find out what they are.

Here is the information I need to put into the computer for your particular printout:

1. Latitude (degrees, minutes, seconds).

2. Longitude (degrees, minutes, seconds).

3. Variation in degrees and whether East or West.

4. What name you wish to appear for location.

You will note from the example for Greenville, the name "From: Greenville, New Hampshire" appears at the top of each page. In my case, I use the coordinates of my house (off the deed to the

FROM: GREENVILLE, NEW HAMPSHIRE

LOCATION	DEG. TRUE	DEG. MAGNETIC	DIST. STAT.	DIST. KILOMTR.	DIST. NAUT.
45. MONT., GREAT FALLS	292	308	1902	3062	1652
46. N.CAR., ASHEVILLE	229	245	802	1290	696
47. N.DAK., BISMARK	289	305	1415	2278	1229
48. N.DAK., FARGO	289	305	1228	1976	1066
49. N.J., NEWARK	211	227	239	385	208
50. N.J., TRENTON	213	229	286	460	248
51. N.MEX., ALBUQUERQUE	264	280	1938	3118	1682
52. N.Y., ALBANY	235	251	125	201	109
53. N.Y., BUFFALO	263	279	358	576	311
54. N.Y., NEW YORK CITY	208	224	239	384	207
55. N.Y., ROCHESTER	265	281	291	468	252
56. NEBR., OMAHA	271	287	1236	1989	1073
57. NEV., LAS VEGAS	272	288	2326	3743	2020
58. NEV., RENO	78	94	2657	4276	2307
59. OHIO, CINCINNATI	248	264	729	1172	633
60. OHIO, CLEVELAND	256	272	525	845	456
61. OHIO, COLUMBUS	250	266	630	1013	547
62. OHIO, TOLEDO	261	277	613	987	533
63. OKLA., OKLAHOMA CITY	256	272	1476	2375	1282
64. ONT., OTTAWA	303	319	224	361	195
65. ONT., TORONTO	272	288	378	608	328
66. ORE., PORTLAND	291	307	2509	4037	2178
67. PA., HARRISBURG	229	245	352	567	306
68. PA., PHILADELPHIA	215	231	311	501	270

fig. 1. An excerpt from the U. S. antenna headings chart plotted on the HAM RADIO magazine office in Greenville, New Hampshire. Antennas can be accurately oriented with an ordinary compass by the magnetic degrees heading.

FROM: GREENVILLE, NEW HAMPSHIRE

BEAM HEADINGS AND DISTANCES

	CALL	LOCATION	DEG. TRUE	DEG. MAGNETIC	DIST. STAT.	DIST. KILMTR.
1.	3A	MONTE CARLO, MONACO (E)	60	76	3808	6128
2.	3VB	TUNIS, TUNISIA (AF)	65	81	4183	6732
3.	3W8	SAIGON, VIETNAM (A)	2	18	8670	13953
4.	4S7	COLOMBO, CEYLON (A)	34	50	8513	13700
5.	4U1	UN, GENEVA, SWITZERLAND (E)	58	74	3662	5893
6.	4U8	UN, TURIN, ITALY (E)	59	75	3759	6050
7.	4W1	SANA, YEMEN (A)	61	77	6676	10744
8.	4X4	TEL AVIV, ISRAEL (A)	56	72	5454	8777
9.	4Z4	TEL AVIV, ISRAEL (A)	56	72	5454	8777
10.	5A	TRIPOLI, LIBYA (AF)	68	84	4471	7195
11.	5B4	NICOSIA, CYPRUS (A)	54	70	5245	8441
12.	5H3	DAR ES SALAAM, TANZANIA (AF)	80	96	7598	12227

fig. 2. An excerpt from the DX beam heading chart.

house, by the way) and so on mine it says: "From: Hoff House, California." I could have said "From: Los Altos Hills, California" just as easily.

You can find your latitude, longitude and magnetic variation from your property deed (the local city hall can look it up *for you), from the* county *surveyor* who has records of all this information, or there are several other things you can do. If you are reasonably close at all to any airport or commercial radio or tv transmitter, they always know exactly what the latitude and longitude are plus the magnetic variation in that area.

You *can* estimate from a *road* map your location with respect to that landmark. If all else fails, give the name of your community, and if it is less than 10,000 population, give the name of some larger nearby community with your approximate distance and direction from there. Again, I can provide you with information just as accurate as you give me to work with.

This should give you all the information you need to get your personalized beam heading printout. Include a self-addressed stamped envelope if you ask any questions that need to be answered, as the computer is located in Texas and all the actual printouts will be mailed from there. Also, I may find this service is too overwhelming to continue, or too expensive to offer for this nominal fee, and may wish to return your money. In any event it shall only be possible to crank out 15-20 of these in one evening, probably, so be patient on getting your copy.

If ordering, provide the following:

1. Latitude.

2. Longitude.

3. Magnetic variation (if you can't find this, I know already for anywhere in the USA to the closest 1°).

4. Name of location you wish to appear.

5. SASE for possible return of money.

6. Include $2 for the printout plus 25c mailing.

7. If sending from outside the United States, mailing costs would be for 2.5 ounces.

Mail to: Irvin M. Hoff
Attn: Beam
12130 Foothill Lane
Los Altos Hills, Calif. 94022

ham radio

the ham notebook

ic power

Too often, when radio amateurs start experimenting with IC packages we stop thinking like amateurs and start thinking by the book. A case in point was my own experience with a dual two-input gate which was used to build a Schmidt trigger. Briefly this is a circuit which will take almost any waveform provided it is above the needed trigger level and convert it into a form which can be used to trigger flip-flops.

The circuit was put together with an input voltage of about three volts, the output on the scope was a very nice square wave. Trying to trigger a series of flip-flops with this output was sheer frustration. Everything was wired properly, the supply voltage was on the money, but the flip-flop triggering was erratic. I finally decided to think like an amateur and measured the *output* of the Schmidt trigger and found that it was about 0.4 volts. I then powered the trigger with my variable voltage supply and found that at six volts dc applied to the IC, the output rose to 0.65 volts and the flip-flops triggered reliably. This circuit has been working very nicely for some months now with six volts, not the 3.6 as recommended. There have been no signs of failing or blowing up. Don't be afraid to think like an amateur!

A. S. Joffe, W3KBM

75A4 hints

To increase the amplitude of the 100 kHz markers on the 75A4, directly substitute a 6BZ6 in place of V1, the 6BA6 calibrator oscillator tube. If a further increase is desired, the oscillator 1-pF coupling capacitor, C5, can be replaced by a 10-pF silver-mica capacitor. The combination of the foregoing will result in a 20-dB calibration signal increase, as read on the S-meter, on the 14 MHz band. This was without any apparent degradation of the frequency stability as read on a General Radio model 1192 frequency counter.

For the convenience of having a front-panel ground of the antenna input circuit for testing, simply bend in the tip of the outside plate on the stator of the antenna trim capacitor, C18. This will short the antenna terminal to ground as the rotor passes over this one point. It means, however, that the trimming capacitor must be rotated from the opposite direction while in use. Full 360° rotation is no longer possible, but this loss is more than justified by the convenience.

M. H. Gonsior, W6VFR

wet basement alarm

Having always been concerned about plumbing leaks in my basement hamshack, I finally whipped up a combination water alarm and shut-off circuit for my well pump. Sensors placed strategically around the basement (near the well pump, washing machine and hot water heater) trigger Q1 into conduction

when a water leak is detected. Q2 is then cut off, releasing the relay which breaks the line to the well pump, preventing the basement from being turned into a swimming pool. When Q2's collector voltage rises, Q1 is latched on via the 470k feedback resistor until the circuit is manually reset. At the same time, Q3 is actually more than enough to operate the relay.

I found that small pieces of pc board worked well as sensors. I used pieces about ½-inch wide and 2-inches long, double sided, with one lead connected to each side.

Al Donkin, W2EMF

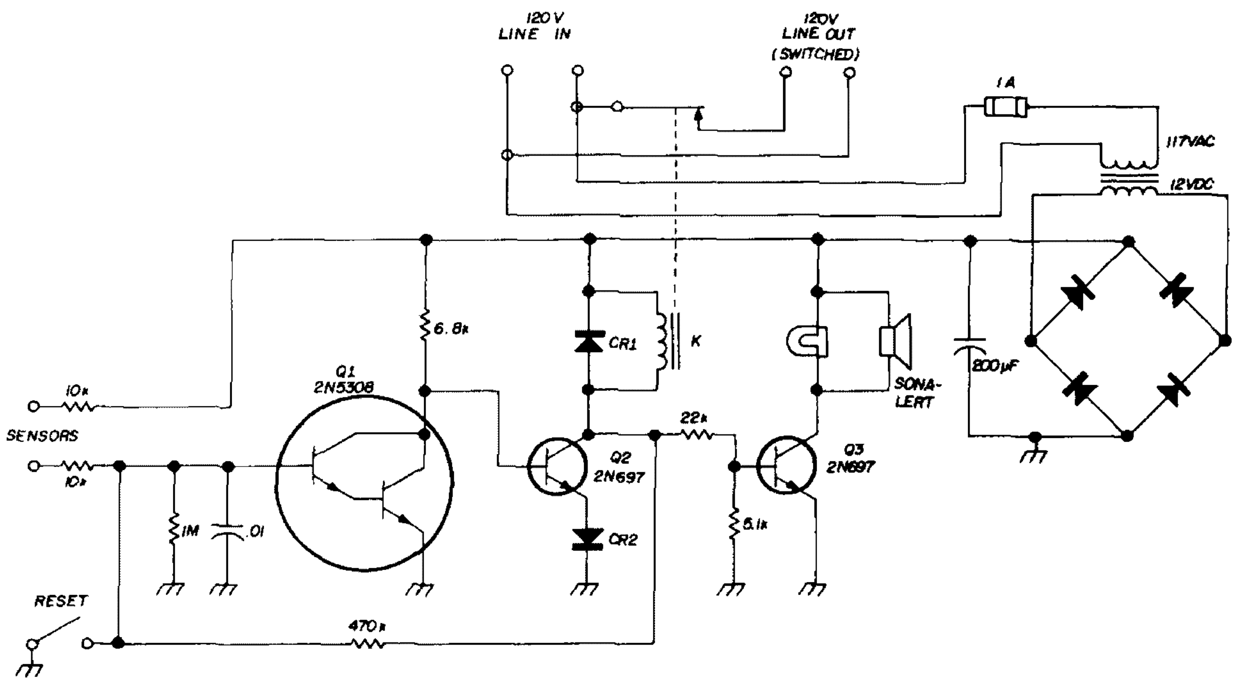


fig. 1. Schematic of the wet basement alarm. The relay is a Magnecraft 88DX3 and the alarm is a Mallory Sonalert.

turned on, sounding the alarm and illuminating an alarm lamp.

Although my installation controls a well pump, a solenoid in the water main supply could be used by the city dweller. Variations on this circuit design may be made to suit the builder's junk box, but a few cautions should be observed. Q1 (2N5308) is a dual transistor internally connected as a Darlington circuit, resulting in high current gain which produces a sensitivity in excess of one megohm for the sensor. The .01-μF capacitor and 10k resistors at the sensor inputs serve as an rf filter to prevent my transmitter from energizing the alarm. The relay I chose is a model with heavy duty (50 A) contacts and a 24-Vdc coil. The power supply produces about 22 volts under load,

s-line spinner knob

S-Line users will find that a small amount of weight added to the receiver tuning knob will materially increase the ease of tuning because of the added inertia of the system. This may be easily accomplished in several ways.

Apply RTV or any other adhesive mixed (such as bathtub caulking) with lead shot, or similar material, to the cavity in the knob. If you put too much weight in the knob, you can remove it with a hobby knife. Alternatively, a metal insert may be fitted into the cavity. For those wishing to splurge, Collins will supply a weighted spinner tuning knob. The part number is 547-1824-003, and the price is ten dollars.

M. H. Gonsior, W6VFR

logic monitor

I call this unit the logic monitor, and it is capable of monitoring logic circuits for dc levels (zero or one logic state) or detecting either positive or negative going pulses. The unit is very simple to make, requiring only a few simple parts and two inexpensive integrated circuits.

The circuit consists basically of an inverter operating into two mono-stable multivibrators. The first monostable operates for positive-going signals and the second one operates for negative-going signals. In detecting negative-going signals, the first monostable acts as an inverter only. Each monostable will operate for its own particular pulse polarity for the length of time set by the R-C time constant of each stage. The time constant of each stage is designed to be long enough to allow the indicator lamp to turn on when a pulse is detected. As the frequency is increased the lamp will glow, but at a reduced brightness.

The logic monitor is invaluable in checking out digital projects, particularly if you do not possess an oscilloscope. This little unit will make short work of checking out a digital keyer, counter or voltmeter. It can determine if flip-flops are toggling or if gates, switching or shift registers are shifting. It is even capable of detecting pulses which many oscilloscopes can not. This is possible since the monitor is designed using medium speed logic classified as DTL – diode transistor logic – which is capable of faster operation than average oscilloscopes.

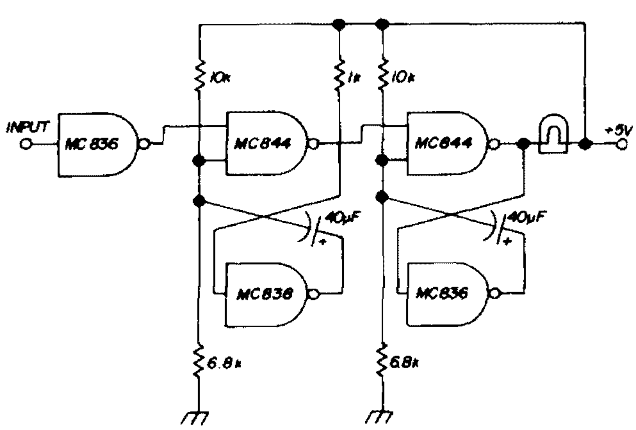


fig. 2. The 844 is a dual 4-input expander with all unused inputs tied to 5 volts with the exception of the expander input.

I believe the logic monitor is a very worthwhile investment. It can be built as a logic probe by installing the unit in a length of tubular plastic or it may simply be contained in a small aluminum box. The unit is designed for five volts and should not be used for logic levels greater than this.

The indicator lamp can be any five-volt lamp which draws 50 mA or less. The monostables are designed using a MC844 which enables the second monostable to also act as a lamp driver, thus reducing the necessity for an additional transistor.

W. L. McGehee, WA5SAF

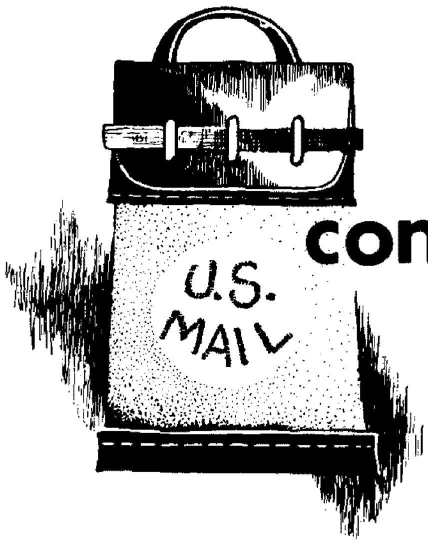
vectorbord tool

Reader's of W6CMQ's article in the August, 1971 *ham radio* might be interested in a commercial tool to accomplish a similar task. Designed for isolating terminal points from a surrounding copper-clad surface, Vector offers two tools to do the job. W6CQM gave many applications for the technique, and Vector points out another – use of inexpensive uninsulated push-in terminals rather than the more expensive insulated standoffs.

The Vector tools are intended for cutting rings through the copper surrounding a prepunched hole. Two sizes are available, the model P116 for ¼-inch diameter circles and the P116A for 3/16-inch diameter circles. They sell for $2.32 and $2.48, respectively. Unlike W6CMQ's version, these commercial tools have a centering pin which limits their use to prepunched or drilled boards. However, to use them with inexpensive, surplus copper-clad board, you would simply have to first drill a small pilot hole for the tool.

The Vector Pad Cutter Tools are available through the Allied catalog and are made to fit most common electric drill chucks.

Douglas Stivison, WA1KWJ



comments

trapatt diodes

Dear HR:

While employed by Boeing Aircraft during the early months of 1970, I was doing research on Trapatt (for TRApped Plasma Avalanche Triggered Transit) and Impatt (for IMPact Avalanche Transit Time) oscillators and amplifiers. Co-workers and I were using diodes developed in the Boeing Silicon Laboratory.

I discovered that most common silicon signal diodes tried would oscillate in the Trapatt mode. In fact, while not performing as well as Boeing diodes, signal diodes did tend to "drop" into the Trapatt mode with less tuning. A fixture was then designed and built which involved very little metal work and cost. A summary of the results obtained with this circuit is given in **table 1.**

In all cases pulse lengths were 0.5 μsec with a pulse frequency of 1 kHz. The pulse frequency probably could have been increased an order of magnitude: pulse lengths *much greater than 1.0 μsec* would probably result in burnout.

The power output and efficiencies are lower than those listed in *"A Second Look"* for October 1971. The frequencies are higher and, in most cases, located in *amateur bands.* My feelings at the time were that CW operation in the Trapatt mode with inexpensive diodes would be most difficult if not impossible. I therefore concentrated my efforts to obtain operation in the 2.3-GHz band and higher – where pulsed operation is allowed. Even with properly designed diodes, the difficulty in achieving good Trapatt results increases rapidly with frequency. This becomes evident when comparing the results obtained with the same FD600 diode at 2.3 GHz and at 3.34 GHz.

The Impatt pumping frequency of the diodes in this circuit was as high as 11.4 GHz in some cases. The circuit may be used with the signal diodes as an Impatt oscillator merely by using a different tuning technique. The circuit did not appear to be as efficient in the Impatt mode as waveguide designs. Outputs were obtainable, however, from S-band to X-band using this circuit.

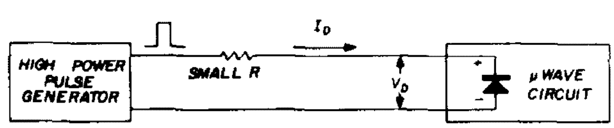


fig. 1. Bias diagram of the Trapatt experimental setup.

Trapatt mode operation is easily identified not only by sudden increase in output power but also by dramatic and exciting increases in diode current and "drop" in diode voltage. A bias diagram will help to describe the phenomenon. The diode is reverse biased.

The pulse amplitude is increased until a certain pulse current is achieved. This varies considerably, but is about 500 mA for most signal diodes. The microwave circuit is then mechanically tuned. When tuning is adjusted such that the diode sees the proper impedances to support the Trapatt mode, the diode voltage V_d will suddenly decrease and the diode current

table 1. Summary of various diode performance in the Trapatt mode.

diode	1N914	1N3064	1N4148	FD333	FD600	FD600
power out (watts) (peak pulse)	3	26	8	22	4	1.2
frequency (GHz)	2.25	1.2	1.2	1.7	2.4	3.34
efficiency (percent)	1.1	10.0	8.5	14.5	4.9	1.8
V_B (volts	120	100	120	200	90	90

suddenly increase. An increase from 200 mA to 1100 mA in diode current was observed with one FD333 diode. More typical values are a voltage decrease and current increase by a factor of two. A sketch of diode voltage, current and rf pulses is shown in **fig. 2.**

A current and voltage change may occur without a subsequent rf output pulse, a phenomenon peculiar to signal diodes. None of the Boeing avalanche diodes tested displayed this mode. Neither types of diodes oscillated at subharmonic frequencies or at Trapatt efficiencies without the current and voltage change.

It is important that the experimenter include in his data reverse capacitance vs voltage plots for each diode tested. The C vs V curve published by the diode manufacturer may be used but is not as useful since C vs V curves are different from diode to diode. C vs V data should be plotted on log-log paper with capacitance on the dependent axis. The slope charactertistics and value should be noted for each diode. Slope is measured on the graph by using a *linear* rule to measure the ratio of rise to run.

An ideal 1- to 3-GHz Trapatt diode

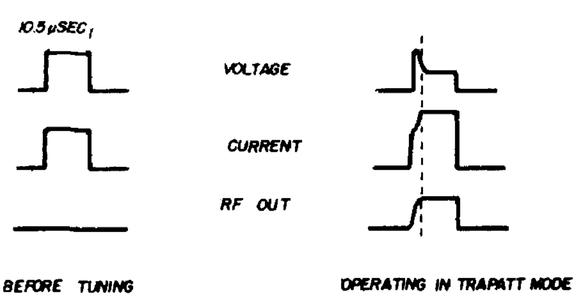


fig. 2. Typical diode voltage, current and rf pulses observed by WA7LNA.

will have a slope value approaching 0.5, the theoretical maximum. Higher breakdown voltages usually signify lower frequency operation. Approximately 100 volts is optimum for 2.0 GHz. A tendency for the slope to decrease suddenly at a reverse voltage 1/3 to 2/3 of V_B is considered desirable; however, the only signal diode found to have this characteristic failed to operate in the Trapatt mode! Many exceptions to these guidelines will be found, but they may serve as a beginning point.

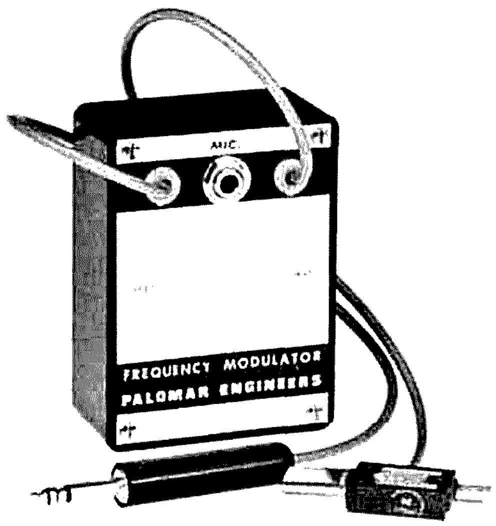
It is my belief that an enterprising amateur, by experimenting with different circuits and inexpensive diodes, could make significant contributions in the field of transit-time microwave devices, even possibly discovering new modes. For example, I am not convinced that all the results obtained in various labs throughout the country qualify to be lumped into one giant Trapatt mode category. It is very likely different labs are calling seemingly similar results operation in the Trapatt mode, when possibly more than one phenomenon is being observed. Why did signal diodes operate with much lower efficiency than Boeing diodes, but tune into the Trapatt mode with greater ease? Why did some 1N4148 diodes with nearly flat C vs V curves operate better than an FD333 with an ideal curve?

I would be most happy to provide assistance to anyone wishing to experiment in this very exciting area. The circuit used in these tests is available to anyone seriously interested in experimentation.

Randall W. Rhea, WA7NLA
920 W. Indian School
Phoenix, Arizona 85013

new products

fm adapter for the Communicators



The Gonset Communicators have been among the most popular two-meter transmitter-receivers of all time. But they are a-m rigs and two meters is now heavily fm.

Palomar Engineers' new *frequency modulator* puts the Communicators on fm without any modification or rewiring. The microphone plugs into the *frequency modulator*. One of the cables coming out of the *modulator* plugs into the Communicator's microphone jack and carries the push-to-talk line, but no audio. A second cable has the crystal plug. It plugs into any one of the Communicator crystal sockets and the crystal plugs into it. A variable capacitance diode in the crystal plug frequency modulates the crystal and the frequency multipliers in the Communicator increase the deviation to about 8.5 kHz at the output frequency.

The *frequency modulator* works with the Communicator I, II, III, IV and GC-105. There are no tuned circuits in the *modulator* so operation is not restricted to the two-meter amateur band alone. Built-in tone burst is available for use with repeaters. The half-second tone burst is keyed by the push-to-talk switch. A carrier frequency adjustment allows the frequency to be set exactly.

Other modulation, such as frequency shift keying, can be applied to the *frequency modulator*. An audio level of 10 mV is required. The frequency response of the *modulator* is 200 to 3500 Hz. Clipping on voice is approximately 7 dB.

The unit is priced at $34.50 postpaid. Built-in tone burst is $10. Communicator model and tone-burst frequency must be specified with order. For more information write to Palomar Engineers, Box 455, Escondido, California 92025 or use *check-off* on page 110.

improved signal one

Computer Measurements, having taken over Signal/One, has announced an improved version of the CX7 transceiver. The new radio, designated the CX7A, incorporates a series of modifications found desirable after years of use of the CX7 in the field. The major modifications include power supply protection to eliminate transients and surge problems which have destroyed sensitive solid-state circuitry in the CX7. The audio passband in both transmit and receive has been broadened to enhance the lower-frequency response. Vox turn-on has been changed to cure the syllable-clipping problems old owners have complained about. TVI and spurious outputs have been reduced by changes to the rf driver circuitry. A few minor changes have also been incorporated in the new CX7A.

Signal/One sells the CX7A for $2195,

More Details? CHECK—OFF Page 110

the same price as the original CX7. Owners of the older model can have all the factory modifications incorporated in their transceiver for $69.50. The entire set of modifications, besides improving overall reliability, is said to give the unit a new sound.

For more information on the new CX7A or on modification for A CX7, write to Signal/One, 1645 West 135th Street, Gardena, California 90249.

rf fet design kit

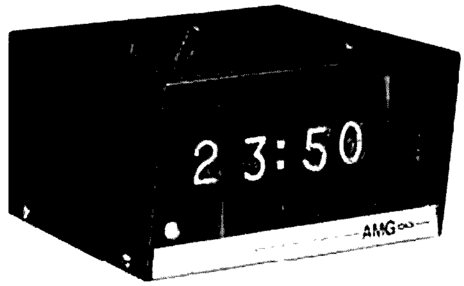
A special rf fet design kit is being offered by Siliconix to help familiarize amateur experimenters and professional designers with the capabilities of rf fets. The kit includes three E300s suitable for fm preamps, two 2N5397s for mobile rigs, two U310s for community antenna television amplifiers, two UT100s for uhf preamps, one 2N5912 for a uhf/vhf mixer, a set of fet design ideas, application notes and data sheets and a copy of the *FET Handbook.*

The regular retail price of the kit would be $55.15, but it is being offered as a package for $19.95 from local Siliconix distributors. It is also available by mail for an additional $1.50 postage and handling fee from Siliconix Incorporated, Attention: Mr. B. Siegal, 2201 Laurelwood Road, Santa Clara, California 95054. More information is available from this address or by using *check-off* on page 110.

worldradio

Wordradio is a new amateur-radio newspaper published every three weeks by Armond Noble, WB6AUH. Completely non-technical and non-political, *Wordradio* covers public service, humanitarian and international aspects of the hobby as well as FCC news. Service nets, ham expeditions, unusual amateurs, amateur connected rescues and mercy missions are all reported in the new newspaper. For a sample copy write to *Worldradio,* 2509 Donner Way, Sacramento, California 95818.

digital clock



Aero-Metric is now offering their new, all solid-state digital clock in both a 12- and a 24-hour version. Time display is in hours and minutes, using bright red neon readout tubes rated at 200,000 hour life (over 22 years). The circuit uses TTL logic and includes 15 integrated circuits, 4 transistors and 7 diodes.

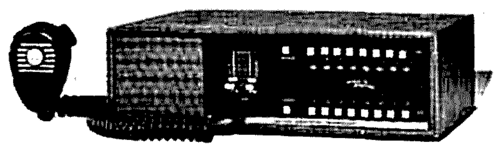
The time base of the logic circuit is taken from the 117-volt power line frequency of 60 Hz so accuracy is based on the United States power grid 60 Hz standard of within 3 seconds per year. The power supply has a built-in, rechargable battery which holds time in the logic circuit during short interval power line failures up to 5 minutes.

The 24-hour clock has a station-identification feature for amateur operators which consists of a bright red neon light which flashes for 30 seconds every 10 minutes. This serves as a reminder to transmit the station call letters as required by FCC regulations. This station-identification feature can be switched off.

The 12-hour clock is $93.00, and the 24-hour clock is $99.00 with standard metal cabinets in a choice of black or gold. A choice of walnut or maple cabinets is available at $9.00 extra. These clocks are unconditionally guaranteed for one year on all parts and labor, under normal use.

Clocks can be purchased direct from the factory with BankAmericard or Master Charge (include card number), C. O. D. ($20.00 down), check or money order. All postage is prepaid except C. O. D. For free literature write to Aero-Metric General, Inc., 155 Franklin Street, Dayton, Ohio 45402 or use *check-off* on page 110.

scanning fm transceiver



Regency Electronics is now marketing their Transcan* base and mobile transceivers for the 2-meter fm band. Units are in production and are being distributed throughout the country.

Regency calls Transcan a new reception concept in fm transceivers. The receiver section of the transceiver scans as many as 8 crystal-controlled channels anywhere in the band. Upon reception of a signal, scanning stops, and the receiver monitors the frequency being used. At the end of the transmission the receiver resumes scanning at the rate of 15 channels per second. Each channel can be quickly programmed in or out of service by the push of a button so the receiver will not be locked onto one frequency if the channel is tied up.

All eight transmit channels are also pushbutton selected. When a transmit button is pushed the receiver stops scanning and locks on the receiver channel paired to the selected transmit frequency. Both transceivers – the HR-2S 117-Vac base transceiver and the HR-2MS 13.6-Vdc mobile transceiver – are American made and are fully solid state.

The receiver boasts 0.35 μV sensitivity for 20-dB quieting, selectivity at 6 dB down of ±16 kHz, 45-dB image rejection and 60-dB spurious rejection. Modulation acceptance is ±15 kHz and audio output is 5 watts into a built-in front panel speaker.

The transmitter runs 15 watts output across the entire 144- to 148-MHz band and has adjustable deviation zero to 15 kHz. Spurious and harmonic emissions are measured at 55 dB or more below the

*Transcan is a trademark of Regency Electronics, Inc.

More Details? CHECK–OFF Page 110

carrier.

The mobile HR-2MS sells for $319, and the base HR-2S sells for $349 including a ceramic push-to-talk microphone and crystals for 146.94 MHz Simplex.

For more information write to Regency Electronics, Inc., 7900 Pendleton Pike, Indianapolis, Indiana 46226 or use *check-off* on page 110.

parts kits

Assortments of the most widely used Centralab electronic components are offered in eight new service kits designed to provide hams, experimenters, service dealers and designers with a well-balanced supply of their component requirements. Briefly described, the new kits are: Kit-10F, Fastatch II controls; -20W, miniature wire-wound controls; -30T, miniature trimmer controls; -50A, axial lead electrolytics; -55P, pc lead electrolytics; -60D, general purpose capacitors; -70H, high voltage capacitors and -100P, packaged electronic circuits.

Each kit is housed in a rugged steel frame cabinet with 15 plastic drawers. Cabinet size is 10 x 8 x 6¼ inches, and the cabinets are portable with convenient handles. The cabinets may be stacked in groups or wall mounted. All kits are supplied completely ready to use with components functionally arranged in drawers by value, type and size. Each drawer is prelabeled clearly showing contents. The latest edition of H. W. Sams "Replacement Control Guide" is included in the 3 control units (Kit-10F, -20W, -30T).

More Details? CHECK—OFF Page 110

More Details? CHECK—OFF Page 110

Complete information on these eight new service kits is available from Centralab distributors, or by writing directly to Distributor Products, Centralab, the Electronics Division of Globe-Union Inc., 5757 North Green Bay Avenue, Milwaukee, Wisconsin 53201, or using *check-off* on page 110.

wire stripper

Radio Shack is offering a new automatic wire stripper and cutter in their line of Archer tools. The new tool strips number 24 to number 12 gauge wire in a second. To operate the wire stripper you insert the wire, squeeze the insulated handles and release. Insulation is removed cleanly and completely without nicking or breaking the wire. A strip gauge guides the wire to the correct portion of the blade and assures a uniform stripped length every time.

The Archer automatic wire stripper and cutter is priced at $4.95 and is available exclusively through Radio Shack's more than 1100 stores or by mail from Allied Radio Shack, 2400 West Washington Boulevard, Chicago, Illinois 60680. More information is available through *check-off* on page 110.

light-emitting diodes

The ham and experimenter can now get four different light-emitting diodes in the Motorola HEP line. The new diodes lend themselves to digital displays, burglar alarms, panel lights, digital clocks, frequency counters and other digital readout circuits in which the low power requirements and long life of the light-emitting diode offer new design possibilities.

The new Motorola LEDs include three visible red diodes (P2000, P2001 and P2003) and one infrared diode (P2002). They are available through any HEP dealer. More information is available from Motorola HEP Semiconductors, Box 20924, Phoenix, Arizona 85001 or use *check-off* on page 110.

More Details? CHECK–OFF Page 110

75 cents

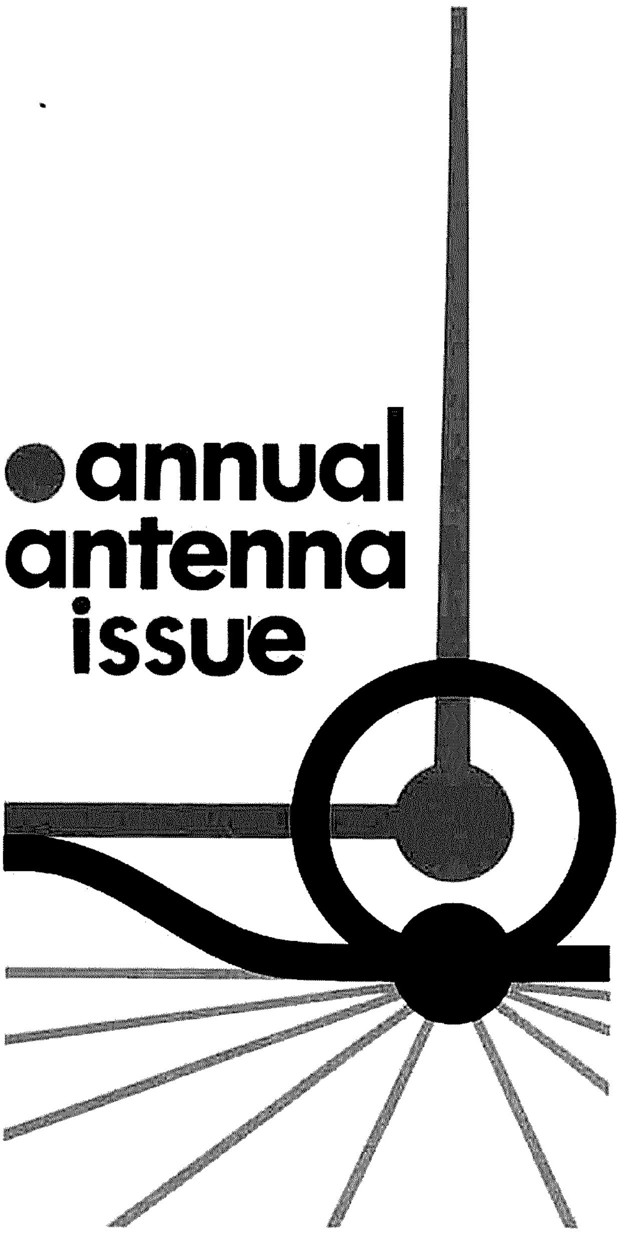
hr

focus on communications technology . . .

ham radio magazine

MAY 1972

annual antenna issue



this month

May, 1972
volume 5, number 5

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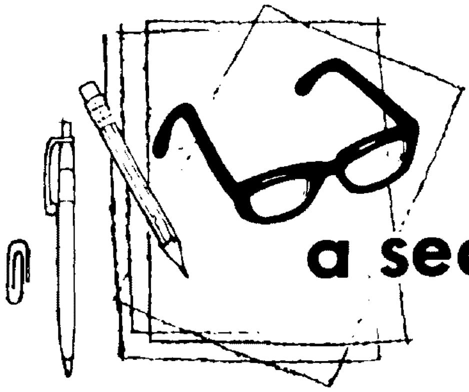


a second look

by jim fisk

Have you been avoiding the exam for the Extra Class license because of the code-speed requirement? Below are some hints that might help you over the hump. It should be remembered, however, that no easy way exists for developing code speed and the confidence so necessary in the examination room. Practice is essential—there's no other way.

Many arguments have been advanced for eliminating the code test for amateur radio licenses. Some of them present a pretty good case, but the fact remains that the FCC requires a *20-wpm* code test for amateur Extra Class, and it looks like this requirement will be around for some time to come. So if you sincerely want to earn that Extra Class ticket you'll have to bite the bullet and improve your code speed.

One method for practicing code is with a tape recorder, but many fellows don't have one. Another method, of course, is by copying other amateurs on the air. This method has disadvantages such as signal fading, inconsistent speed and character spacing, and irregular on-the-air time of stations you're trying to copy.

The code-practice transmissions by ARRL's W1AW are excellent sources of perfect, tape-controlled code, which are available to anyone having a shortwave receiver. But here again problems exist. W1AW is limited to the maximum legal power input of 1 kW for amateur stations and is subject to the caprices of propagation conditions, not to mention interference, intentional or otherwise, by other amateur stations.

If you own a general-coverage receiver, you can augment W1AW's code-practice transmissions by copying commercial coastal stations, which are available around the clock in the high-frequency marine bands of 4, 6, 8, 12, and 16 MHz. These stations handle traffic for ships and are manned by top-notch operators whose job is to get the message through. Therefore, code transmissions from the coastal marine stations are tailored to fit prevailing conditions existing in the radio circuit. This means a reliable source of code practice is available with excellent code at speeds between 10 and 35 wpm. Another advantage is that these stations operate on clear channels, and their frequencies in the hf bands assure consistent communications during most propagation conditions.

The following table lists some of the public-service coastal stations in the 8-MHz marine band that provide excellent code practice. Several stations (WNU for example) transmit taped press items at a steady, interference-free 25 wpm—wonderful for building up your code speed.

Station	frequency (MHz)	power (kW)	location
WSL	8.514	2.5	New York, N. Y.
WSL	8.658	10.0	
WCC	8.586	10.0	Chatham, Mass.
WNU	8.570	2.0	New Orleans, La.
WSC	8.610	20.0	Tuckerton, N. J.
WSC	8.686	2.0	
WPA	8.550	3.0	Pt. Arthur, Tex.
KOK	8.590	10.0	Los Angeles, Cal.
KPH	8.618	10.0	San Francisco, Cal.
KFS	8.558	10.0	San Francisco, Cal.

These are only a few of the stations available as code-practice sources for those who really want to earn the coveted Extra Class amateur license.

Will I see you down on the "basement frequencies?" I hope so.

Jim Fisk, W1DTY
editor

three-band groundplane

Theory and design of a simple three-band groundplane featuring optimized 20-meter DX performance and desktop bandswitching

Peter Brekken, LA1EI, Vegmester Kroghs gt. 24B, 7000 Trondheim, Norway

Vertical antennas are popular *among* those of us who have little space to erect an antenna. The vertical antenna has certain desirable features – and unfortunately, some less desirable ones also. Among its good points, it is usually non-directional, yet it can be made to give some gain over a horizontal half-wave dipole. Vertical elements can be combined in an array to give a desired gain or radiation pattern, just as you can do with horizontal elements.

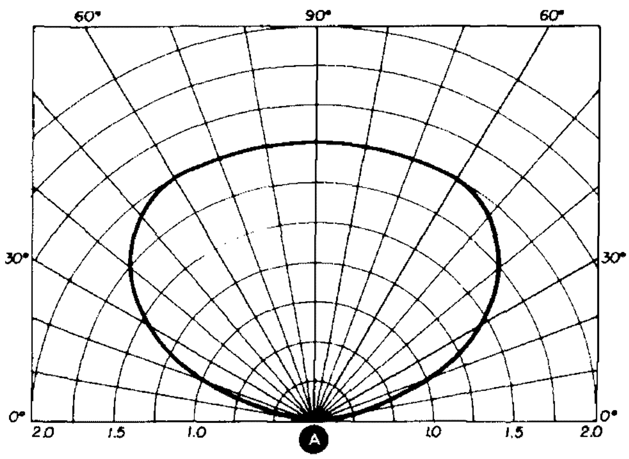
radiation angle

The radiation from wire antennas is polarized in the direction of the wire; thus the vertical antenna gives vertical polarization with the major part of the radiation occurring at a much lower angle than the radiation from a horizontal dipole. Fig. 1 shows the vertical-plane radiation pattern for a horizontal dipole at various heights above ground.* Because of the reflections from the ground, these patterns are very different from the theoretical free-space pattern shown in fig. 2. The patterns are independent of the length of the dipole, but as the diagrams show, are very dependent on the height above ground.

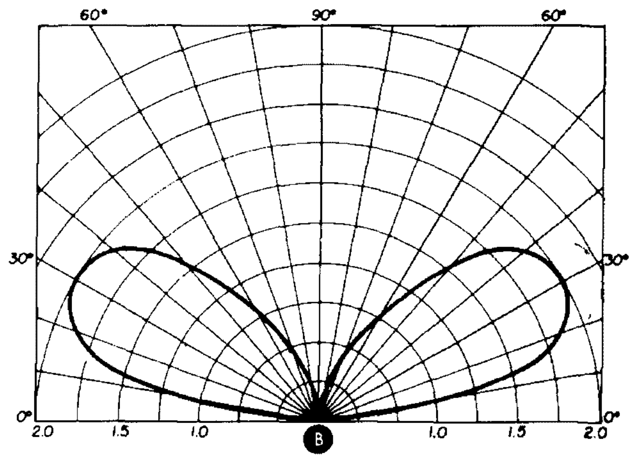
Thirty to forty feet would frequently be considered a quite acceptable height for a halfwave dipole. On 40 meters this equals about ¼-wavelength, and we can

The ARRL Antenna Handbook* contains a far more complete collection of radiation patterns for horizontal dipoles at heights from 1/8 to 2 wavelengths. This book is highly recommended to those interested in further study, construction or experimentation with antennas. **editor.

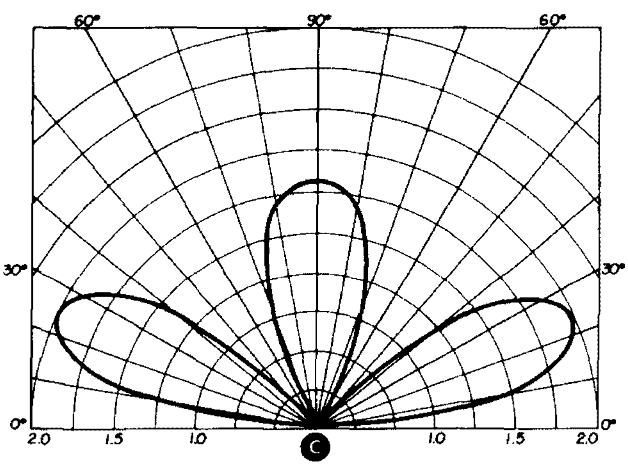
apply **fig. 1A.** From the figure we see that the major part of our precious few watts of rf are radiated straight upwards – into the empty space. Since we are usually not interested in space communication on 40 meters, this will not make a very efficient antenna. On 20 meters the same height equals ½-wavelength, and from **fig. 1B** we can see that the major radiation now occurs at an angle of approximately 30°. At 10 meters, when the height is 1- to 1½-wavelengths, the radiation splits into more lobes, and we still have an appreciable amount of the radiation at high angles.



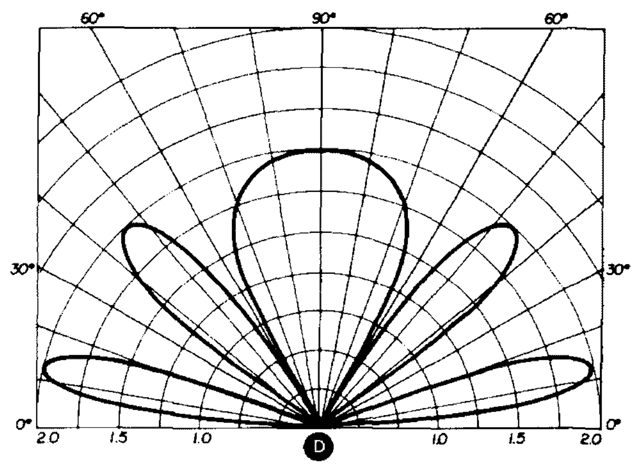
Antenna at ¼-wavelength. Notice large amount of power wasted in straight vertical radiation.



Antenna at ½-wavelength. Note how the majority of power is transmitted at an angle of 30°.



Antenna at 5/8-wavelength. At least two-thirds of the available power goes into useful, low angle radiation — but not so low as to be totally absorbed by the earth.



Antenna at 1¼-wavelength. Notice how power is again being wasted in vertical radiation which the ionosphere will not reflect back to earth.

fig. 1. Vertical plane radiation patterns from horizontal dipoles at different heights about a perfectly reflecting ground.

propagation

It is known that high-frequency radio waves (3 to 30 MHz) are to some extent reflected from the layers of ionized air in the stratosphere, mainly the so-called F2 layer which exists at an altitude of about 150 to 250 miles. When the reflected wave reaches the earth again, it can be received. In this way, distances up to 2500 miles can be covered in one hop. Usually

the wave that reaches the earth is reflected upwards again, and can, under good conditions, make at least three or four hops which will cover distances up to 10,000 miles or more.

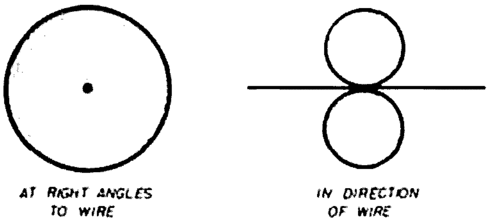


fig. 2. Free-space radiation patterns for a ½-wavelength dipole. Compare these patterns with the patterns of fig. 1 to see the effects of ground reflection.

During each hop the wave is attenuated a certain amount. The wave that covers the distance in the smallest possible number of hops, therefore, will give the best signal strength at the receiver. It is easy to see that the antenna which has the lowest radiation angle will cover the greatest distance per hop and consequently should be the best DX antenna. From fig. 3 we can see that some of the vertical antennas have their maximum radiation at angles below 10° which is significantly less than for the horizontal dipoles we have studied.[1, 2]

The properties of an antenna over flat conducting earth can be evaluated by considering the earth as a mirror. The obvious effect of a mirror is to create an image, and the antenna and its image are symmetrical with respect to the reflecting surface. The resulting radiation can be found by considering the antenna *and* its image as the radiating system. Thus a ¼-wavelength vertical antenna with its image added equals a vertical ½-wavelength dipole. Then it is easy to imagine that the ¼-wavelength vertical should have the same gain as the half-wave dipole. A 5/8-wavelength vertical corresponds to an extended zepp and should give about 3-dB gain over a half-wave dipole.

efficiency

Up to now we have not considered that earth has only a finite conductivity. Our mirror, therefore, is not perfectly reflecting. Some of the energy radiated at the very lowest angles (below 5° or so) will be lost and only used to heat the soil.

One can define a factor of efficiency, η, for an antenna as for other electrical apparatus.

$$\eta = \frac{\text{radiated power}}{\text{input power}} 100\% \approx$$

$$\frac{R_r I^2}{(R_r + R_{loss})I^2} 100\%$$

$$\eta = \frac{R_r}{R_r + R_{loss}} 100\%$$

Here R_r is the radiation resistance and R_{loss} is the equivalent series loss-resistance.

The half-wave dipole has a radiation resistance R_r of about 72 ohms. The loss resistance R_{loss} is simply the resistance of the antenna wire, which is at most a few ohms. The efficiency, therefore, will be close to 100%.

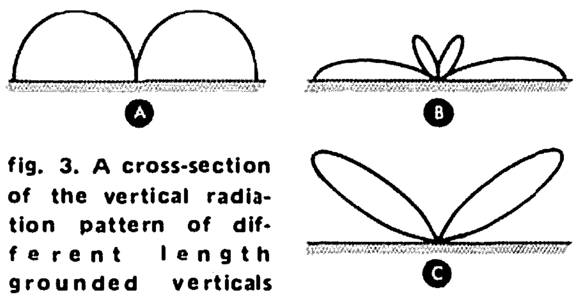


fig. 3. A cross-section of the vertical radiation pattern of different length grounded verticals over perfectly conducting ground. (A) Antenna length is ¾-wavelength. (B) Antenna length is 5/8-wavelength. Notice the majority of power going into low-angle radiation. (C) Antenna length is one full wavelength.

We have considered the vertical antenna as one half of a dipole. As you might expect, its radiation resistance is one half of the resistance of the corresponding dipole. The radiation resistance of a ¼-wavelength vertical is about 36 ohms, and a 5/8-wavelength vertical has a radiation resistance in the vicinity of 70 ohms. The loss resistance depends on the conductivity of the soil and is difficult to calculate. It varies with frequency and other factors and can typically be in the order of some tens of ohms.

In order to achieve a reasonably high efficiency, we see from our formula for η that it is important to keep the ground losses small and the radiation resistance high. The radiation resistance can be increased by increasing the antenna length. The loss resistance is somewhat tricky to minimize. Wet soil is a reasonably good conductor, so the wettest point in the garden would seem to be the best point to place a vertical. A well known trick used at broadcast stations is to install a ground net consisting of as long and as many radials as possible, either on the ground or buried a few inches below the surface. Fifteen radials at least 1/8-wavelength long should give a worthwhile improvement.

the groundplane

The vertical does not necessarily have to be placed on the ground. The groundplane antennas have an artificial groundplane usually constructed from radials. The whole structure can then be placed on the roof or at any convenient location so as to get clear of obstacles that might otherwise be screening the radiation. To get the best results there should theoretically be an infinite number of infinitely long radials. However, many people are getting excellent results with the ¼-wavelength vertical with two to four radials each ¼-wavelength long. If the groundplane is at a reasonable height (¼-wavelength or more) you can expect good results from four radials.

The current and voltage distribution on the radials can be considered to be approximately sinusoidal. The current is maximum at the feedpoint and has its first null a ¼-wavelength away from the feedpoint. We do not need any wire to carry the current if it is zero so we might as well cut the radial here. To rf, the ¼-wavelength radials will look like they are of infinite length and have the loss resistance of a ¼-wavelength wire. It is clear, therefore, that the length of the radials is equally important to good operation as is the length of the vertical element.

The radiation resistance of the ¼-wavelength vertical, about 36 ohms, does not give a perfect match to the usual 50- or 75-ohm coaxial cable. Slanting the radials downwards increases the radiation resistance somewhat so that at an angle of approximately 45° a reasonably good match to 50-ohm coax is possible. This is usually very practical when the antenna is mounted on a roof or at the top of a pole.

From what was said before, we know that another way of increasing the resistance would be to increase the antenna length. Increasing it to 5/8-wavelength gives an excellent match to 75-ohm coax. At the same time we get a 3-dB gain over a half-wave dipole. The antenna is not naturally resonant however, so in addition to the ohmic radiation resistance of 75 ohms, the feedpoint impedance has a capacitive reactance of 250 ohms. This reactance can be tuned out by a series coil.

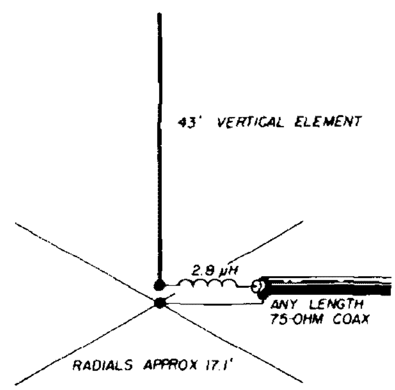


fig. 4. 5/8-wavelength vertical for 20 meters.

a DX antenna

Let us try and design a 5/8-wavelength antenna for the 20-meter band. The length will be about 43 feet, but this is not very critical. We must use a series coil:

$$L = \frac{X}{2\pi f} = \frac{250}{2\pi(14.3)} = 2.8\,\mu H$$

This would make an excellent one-band DX antenna (**fig.** 4).

multiband designs

Many of us would like to use the same

antenna for more than one band. The antenna just described would not work well on 10 and 15 meters. The radiation at the higher angles increases considerably as the antenna length exceeds one wavelength. It could presumably be put to good use at 15 meters, but this is not considered here.

On the lower bands, however, we can get low-angle radiation. If we try to use our 20-meter groundplane on 40 meters, we get a length of about 0.32 wavelength. The gain is about the same as for a half-wave dipole, but with a lower radiation angle. The feedpoint impedance shows a *radiation resistance of about* 140 ohms and an inductive reactance of 260 ohms. To tune out the reactance we can use a series capacitor of 88 pF. If we use 75-ohm coax the swr will be close to 2:1. This gives a loss of about 0.1 dB in 100 feet of coax, which is of no consequence. However, some transmitters will not tolerate this standing-wave ratio.

On 80 meters the length is 0.16 wavelength. The gain is about 0.5 dB down compared to the halfwave dipole. Radiation resistance is low, about 8 ohms, and the reactive part of the feedpoint impedance is about 280 ohms, capacitive. This necessitates a series coil of 12.7 μH. The swr is now very high, 10:1, and with

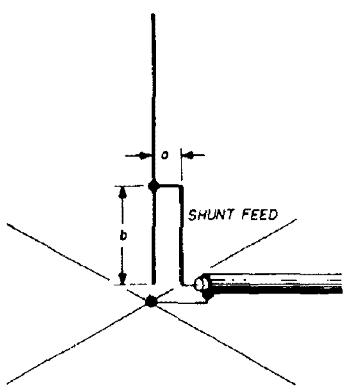


fig. 5. Vertical antenna with gamma match. Suggested dimensions for using the 43-foot antenna on 80 meters would be, a=5", b=10'.

100 feet of 75-ohm cable, there is a loss of around 1.2 dB. The same length of 50-ohm cable (swr = 6:1) gives a loss of 0.5 dB.

RG8U and many other types of cable have a guaranteed power handling capability of 2000 watts. The power rating is usually limited by the maximum allowable voltage that the line will withstand. Standing waves give increased voltages so that the maximum allowable power is decreased proportionally to the standing-wave ratio. An swr of 6:1 will reduce the maximum power to 1/6, or about 350 watts.

matching the antenna

Some sort of matching device would

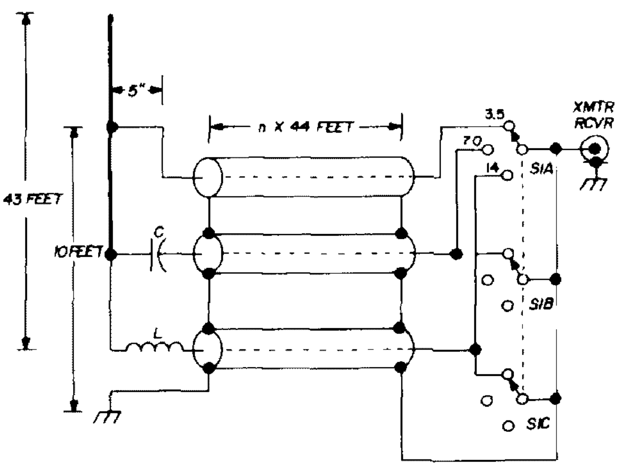


fig. 6. The completed three-band vertical. See table 1 for impedances.

be welcome. A toroidal transformer connected between antenna and feeder could be used.[3] A matching device that would seem to be ideally suited in this case is the shunt-feed or gamma match illustrated in **fig. 5.** The gamma match can give an impedance step up of up to six times. In addition, it exhibits an inductive reactance which usually is unwanted and troublesome. In our case, however, we can use the inductive reactance to cancel the capacitive reactance of the antenna itself, and at the same time obtain a better impedance match to 50-ohm coax. If there are standing waves on a feeder, you know that the feeder will act as an impedance transformer. By using a series coil, capacitor and gamma match, you can make the feedpoint impedance ohmic. If the impedance presented at the transmitter end of the feeder also is to be ohmic, we will have to

table 1. Feedline impedance of the three-band vertical of fig. 6. Impedances are given for feedlines which are even and odd multiples of 44 feet.

band (MHz)	feedline impedance (ohms) odd	even
3.7	25	100
7.0	140	34
14.0	75	75

use a feeder which is a multiple of an electrical quarter wavelength. In coax cable, a quarter wavelength at 80 meters (3.7 MHz) is approximately 44 feet.

If you use this antenna on 80, 40 and 20 meters with a single feeder, you must use a relay (or a switch) at the base of the antenna to switch in the right matching device for each band. Another possibility is to use separate feeders for each band and perform the switching by a simple rotary switch in the shack.

A suggested layout using the latter principle is shown in fig. 6 for use with feeder lengths that are an odd multiple of 44 feet. The feeders that are not in use are shorted by the switch to minimize the influence from the coil and capacitor for the other bands. If even multiples of 44 feet are used as feeder lengths, you can omit the shorting switches S1B and S1C.

All component values are calculated using approximate methods (especially the dimensions of the 80-meter gamma match), and some cut and try may be necessary to obtain the best swr. Both the step-up ratio and the capacitive reactance of the gamma match increase as the gamma section is made longer. They do not increase in the same proportion; as the length is increased the increase in reactance will be faster than the increase in step-up ratio.

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2. W. L. Weeks, "Antenna Engineering," McGraw-Hill, New York, 1968.
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ham radio

nine-element colinear antenna for two meters

The colinear antenna is becoming popular for vhf work — this version has appreciable gain and features some novel construction ideas

Milton Ash, W6RJO

The vertically polarized colinear antenna described in this article is constructed entirely of lightweight aluminum tubing including dipole elements, stubs, and mast. The stubs are integral with the dipoles, being of the same type tubing. This design is in contrast to that of most colinear antennas described in the literature, which use wire elements and stubs supported by nonconductive structures.

The antenna uses nine dipoles, with the metal mast spaced ¼ wavelength from the dipoles. Experimental data indicates that this array provides about 9-dB gain over an isotropic radiator in the horizontal direction. The metal mast apparently acts as a set of individual reflectors for each dipole.

Included are data on construction, tuning, adjustment, and a theoretical discussion on radiation patterns. Also presented are some test results from experiments using a 1/9-scale model antenna to verify certain assumptions.

As with all amateur antenna designs, the data supplied here is subject to modification based on a rigorous theoretical analysis and tests conducted in a laboratory environment. However, the subject is presented as an amateur antenna design, with supporting data based on amateur experiments. The antenna performs as intended, which was

the objective of the entire project.

evolution of the colinear

The colinear antenna evolved from the early Franklin array of the 1920's. The Franklin used wire elements, with or without parasitic reflectors, and was suspended from hundred-foot-high towers. It was originally used for commercial point-to-point long-haul circuits on medium-high frequencies. As higher frequencies came into use, scaled-down versions were used for airborne search radars. As an example, a WW II X-band radar antenna, consisting of some 200 colinear dipoles, was mounted on one side of a horizontally-oriented waveguide. The individual feeds consisted of one leg of each dipole protruding into the guide. A bellows arrangement mechanically oscillated the opposite side of the guide, which varied the phasing of the feed system. The entire array was mounted inside an airfoil under the fuselage of the B-29 bomber. This configuration yielded a windshield-wiper-like oscillating scan, about 60 degrees wide, for scope viewing. The plane of the resulting very thin beaver-tail-shaped beam was perpendicular to the aircraft wings.

two-meter colinear

For the two-meter amateur band I wanted a vertically polarized antenna with the following features:

1. No moving parts, thus avoiding rotator problems.

2. Inexpensive, readily available materials.

3. Appreciable gain, with a pattern that would cover the local Los Angeles area, which is bounded by the sea to the West.

The answer appeared to be a multi-element colinear made of scrap aluminum, mast and all. One 35-40 foot 2½-inch-diameter mast and six 12-foot sections of 3/8-inch tubing (i. e., electrical conduit) are the main items to be acquired.

construction

As shown in fig. 1, the antenna is made of ½-wave dipoles separated by ¼-wave stubs. The dummy stubs, as explained later, have no effect on the antenna electrical characteristics, but are

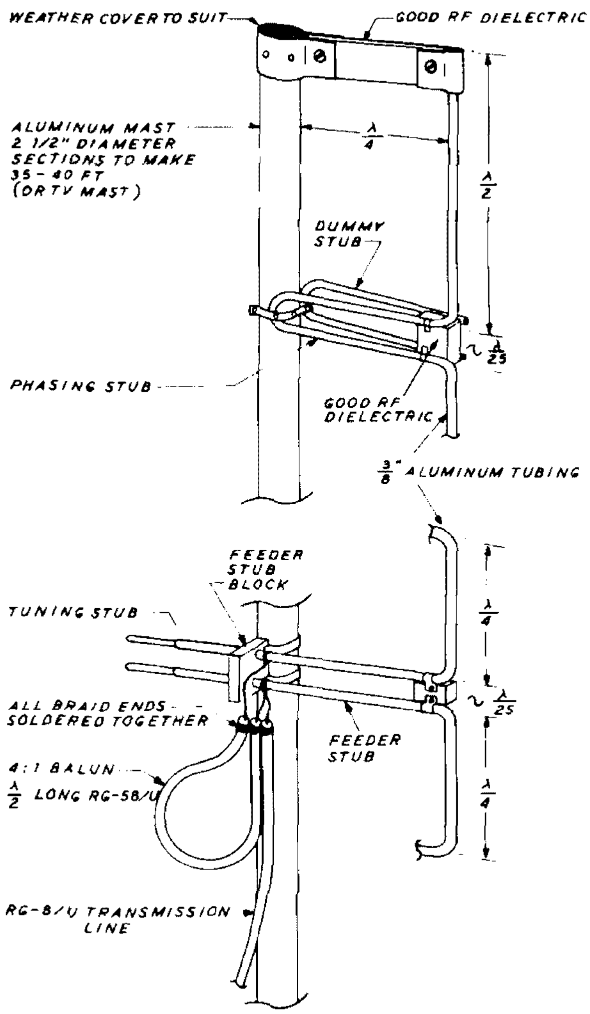
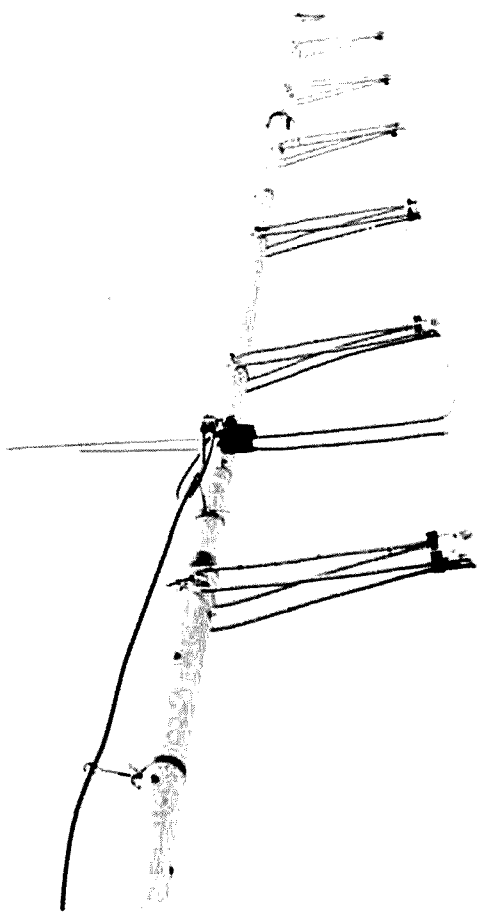


fig. 1. Construction details of the 9-element colinear antenna.

included for added support. Stubs and dipoles are formed from 12-foot pieces of 3/8-inch OD tubing connected together. Nominal lengths of the dipoles and stubs are respectively 38 and 19 inches. Handbook formulas were used so that you may calculate dimensions to suit a particular portion of the band desired. The stubs are nominally 3½ inches wide (1/25 wavelength).

Each bent length of tubing is connected to the next with short pieces of aluminum tubing secured with sheet-

metal screws. In its final form, the antenna will consist of two long lengths of tubing. In my case, one length contained five and one half ½-wave dipoles, five ¼-wave stubs, plus half of the feeder stub. The other length consisted of three and one half ½-wave dipoles, three ¼-wave stubs, plus the other half of the feeder stub.



Feed arrangement and stub detail. Plastic box houses balun and feed connections.

The antenna could be built symmetrically, each section consisting of two identical lengths of four and one half ½-wave dipoles four ¼-wave stubs, plus one-half of the feeder stub. However, I built my antenna as shown to fit my installation requirements. A colinear antenna may be fed by splitting any of its dipoles into ¼-wave elements with the feeder stub, or it can be so fed between any of its dipoles.

tube bending

Aluminum tubing may be bent in several ways. You can carefully heat the material while bending it around a mandrel, or the tube may be filled with sand to provide "body" while bending. I found that the 12-foot lengths could be bent by hand without using a mandrel. Merely bend slowly and carefully to avoid kinks. Make some practice bends with scrap tubing to determine where the tubing starts to kink. You can then determine the optimum bending radius by feel, assuming an average "wrist strength." The bending radii do not have to be as small as shown. With practice, the tube-bending job can be done without special tooling. The "wrist-strength" criteria, together with care and practice, should yield stubs about 3 to 3½ inches outside width. Soft-drawn aircraft-type tubing works well. Aluminum electrical conduit, which may be obtained from mail-order houses, also works well.

dummy stubs

A novel feature of this antenna is the use of dummy ¼-wave stubs for added structural support. Because the open end of a ¼-wave stub represents an infinite impedance, dummy stubs may be added in this manner without impairing antenna performance. The antenna is supported by the phasing stubs and dummy stubs, which are fastened to the aluminum mast using scrap aluminum straps and sheet-metal screws. A good electrical rf connection should be made between the stubs and the mast. This is achieved in part by using many sheet metal screws while remembering that every screw hole weakens tubing. Use all-aluminum hardware wherever possible to avoid electrolysis problems due to dissimilar metal contact.

Note also that pieces of scrap teflon are used for additional bracing at the high-impedance ends of the stubs, at the feeder-stub block, and at the mast supports for the two extreme dipole ends for added stability. Any other good dielectric material can be used, such as polystyrene.

The mast base rests in a Christmas-tree support weighted with a few bricks. However, the principal support comes

from a homemade scrap sheet-iron clamp securing the mast to the eave siding (see photo). Inside the attic are two steel straps bolted through the siding to the clamp. The straps are L-shaped and are bolted to an attic floor joist. Three lightweight nylon rope guys provide addi-

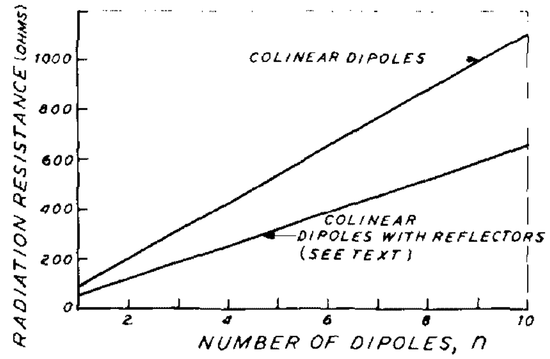


fig. 2. Radiation resistance of a colinear antenna as a function of the number of dipoles.

tional support against the wind. The guys are spaced about 120 degrees apart and are attached two-thirds of the way up the mast.

feeding and tuning

Feeding and tuning procedures are straightforward. Since it was desired to feed this antenna with coax, a balun was needed to match the balanced feeder stub to the coax cable. As depicted in **fig. 2,** the theoretical radiation resistance of a nine-dipole colinear without reflectors is about 1000 ohms. However, an assumption is made (supported by experimental verification of the calculated radiation patterns, as discussed below) that each portion of the aluminum mast acts somewhat as a reflector for its corresponding dipole, so that the antenna can be construed as consisting of nine colinear dipoles with parasitic reflectors. The mast reflectors are spaced ¼-wave from their respective dipoles, and are assumed to be of equal length. **Fig. 3B** shows that a parasitic element of approximately the same length as a dipole and spaced ¼-wave away reduces the dipole input resistance to about 52 ohms. Assuming that this resistance increases with the number of dipoles, it is seen from the lower curve of fig. **2** (drawn to correspond to this trend), that the radiation resistance of nine colinear dipoles with reflectors is about 600 ohms. A 4:1 step-up balun in parallel with the tuning stubs is used to match 52- or 72-ohm coax.

The antenna was tuned by first inserting an swr bridge as close as possible to the feed point. The coax was disconnected about 18 inches from the feed point for this purpose. Two pieces of aluminum tubing about 18 inches long were inserted into the feeder-stub tubing ends to provide a sliding fit. The tuning stubs were then adjusted for minimum swr.

bandwidth

This antenna is sufficiently broadband so that, even though its minimum swr occurs at 147 MHz (**fig. 3A**), the swr is below 2:1 over the entire two-meter band. The stubs and dipoles could be made slightly longer than 19 and 38 inches, respectively, to move the minimum swr point lower in frequency. To

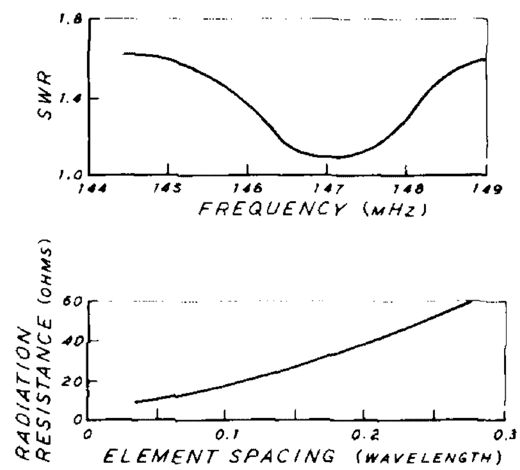


fig. 3. Measured swr of the 9-element colinear over the 2-meter band (upper curve). Lower curve shows radiation resistance of a dipole with parasitic element for various spacings (from reference 2).

calculate such small changes in length, note that the percent increase in required dipole and stub lengths is essentially equal to the percent frequency decrease.

From the photo, it is seen that the tuning-stub lengths are unequal, indicat-

ing asymmetry in the antenna currents about the feedpoint. To achieve symmetry, the fifth dipole should have been split by the feeder stub. Instead, the fourth dipole from the bottom was split to attain a convenient height for tuning from a stepladder. Asymmetrical feed, if not exaggerated (such as feeding from an extreme-end dipole), generally results in only second-order effects on the radiation patterns, including a slight beam uptilt out of horizontal in this case.

radiation patterns and gain

The curves in **fig. 4**, which represent polar radiation patterns in free space, show what may be expected (theoretically) as phased colinear dipoles are added to a simple dipole antenna, also in free space. It is seen that the addition of phased dipoles causes the radiation pattern to exhibit a flatter shape. **Fig. 5** shows the relationship between maximum gain and the half-power beamwidth, as a function of the number of added dipoles, over an *isotropic radiator*. Note that the gain of a single dipole (**fig. 5**) is referenced to approximately 2 dB, which is generally accepted as its gain over an isotropic radiator.

A comparison of power gain in dB may be obtained from the curves of **fig. 4D** if the numerical value of each point of the curves is converted to its logarithm. The power gain, therefore, of a 9-element colinear array is about 8 dB over that of a single dipole, or 10 dB over an isotropic radiator (see **fig. 5**).

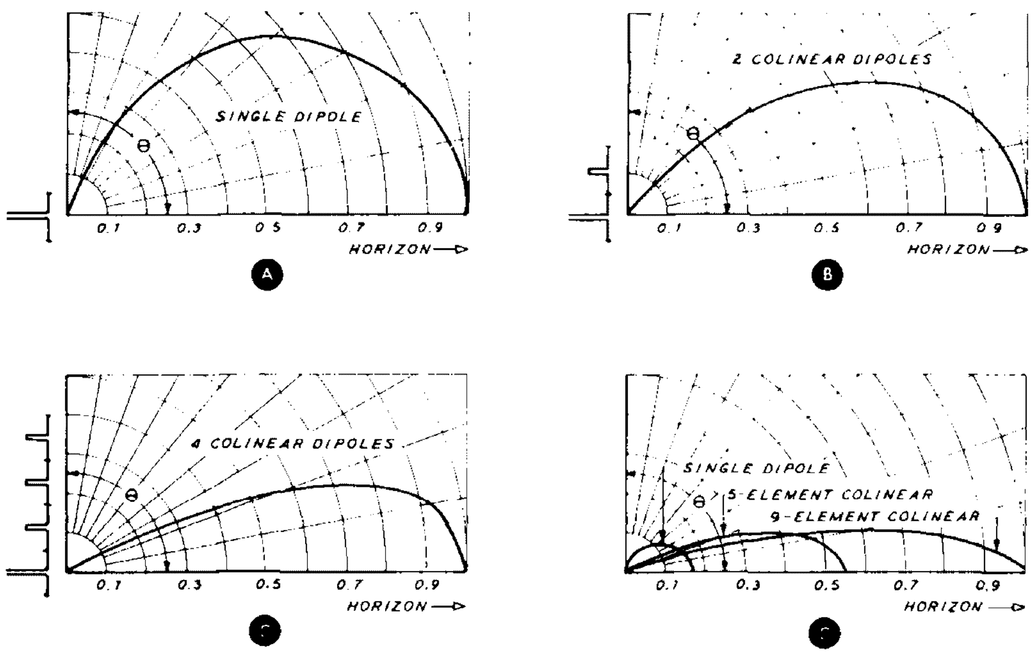


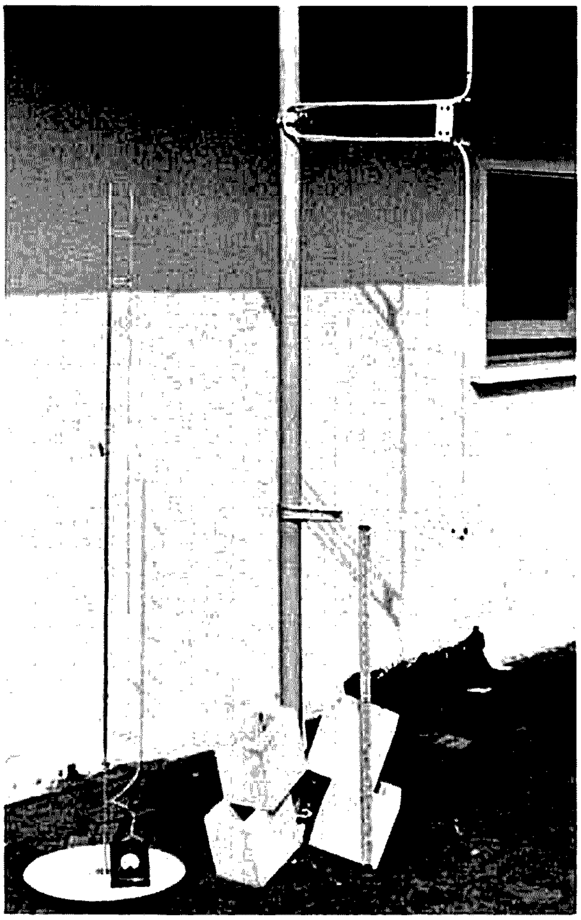
fig. 4. Free-space polar radiation patterns of colinear antennas without reflectors.

The curves in **fig. 4** are graphical expressions of the magnitude of the free-space polar-field radiation patterns of colinear antennas without reflectors:[1]

$$F_n(\theta) = \frac{\sin(90^\circ\, n \cos\theta)}{n \sin(90^\circ \cos\theta)} \times \frac{\cos(90^\circ \cos\theta)}{\sin\theta}$$

where n is the number of dipoles in the array (each dipole is assumed driven by a separate exciter, all of equal power, which is approximated by our single-dipole feed and the phasing-stub arrangement). The angle θ is the zenith angle, measured from directly overhead of the antenna to the horizon for a vertically polarized colinear.*

*The number of minor lobes (not shown) of the polar patterns of **fig. 4** is given by the smallest integer nearest the value (n-1)/2, where n is the number of dipoles. For example, 8 and 9 dipoles would have 3 and 4 minor lobes respectively.



At left, the 1296-MHz antenna used for measurements. Bottom portion of the 2-meter colinear is at right with the yardstick included for scale. Mechanical mounting strength mainly comes from a clamp under the house eaves rather than the simple Christmas tree stand shown here.

The maximum gain in dB, in the horizontal direction, is[1]

$$G_{max} = 10(\log_{10}120 - \log_{10}R/n + \log_{10}n)$$

where R is the radiation resistance. As the number of dipoles, n, is increased beyond three or four, R/n, the array radiation resistance, approaches a constant value of about 110 ohms.[1] Thereafter, the gain becomes proportional to the log of the number of dipoles, so that doubling the number of dipoles gives only a three-dB increase in gain. For example, the gain of a nine-element colinear is about 8 dB over a single dipole. Doubling to 18 dipoles, perhaps by stacking, would yield but 11 dB gain. This logarithmic behavior of gain with the number of elements generally holds for most antenna arrays including quads and Yagis.

azimuthal patterns

The vertically polarized colinear without reflectors has an omnidirectional radiation pattern in the azimuthal plane (plan view of antenna) in free space exactly like a single dipole. With a metal mast, however, the colinear is construed as nine colinear dipoles with reflectors, as discussed earlier. Then, looking down on top of this antenna, the azimuthal radiation pattern is, to good approximation, similar to a single half-wave dipole with an equal-length reflector, spaced ¼ wavelength from the dipole. Such a configuration yields a minor lobe opposite in direction to the major lobe. The minor lobe is 3-6 dB below the major lobe, which is confirmed by experimental data, as shown below.

experimental data

To verify that the metal mast does give a single minor azimuthal lobe opposite in direction to the major lobe, with a front-to-back ratio of 3-6 dB, a 1/9-scale antenna model was built to simulate the two meter colinear patterns. The scale-model antenna was cut for 1285 MHz. It was made of no. 14 tinned wire with a 3/8-inch copper tubing mast. The experimental antenna patterns were obtained using an AN/APX-6 converted surplus IFF transponder transmitter.

The resulting experimentally deter-

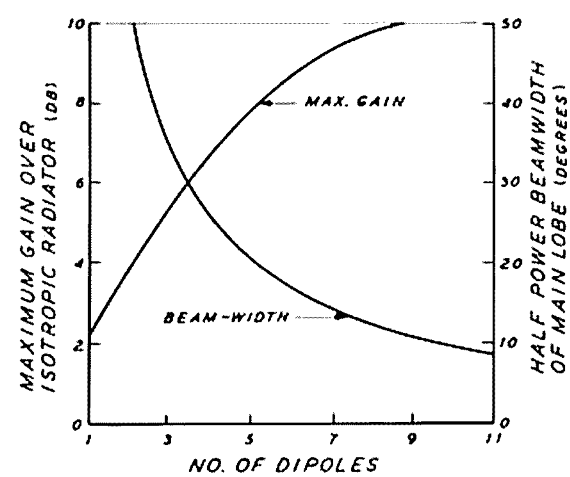


fig. 5. Relationship of maximum gain and half-power beamwidth to the number of colinear dipoles in the array.

mined azimuthal pattern is plotted in fig. 6 and does indeed show the minor lobe opposite in direction to the major lobe, with the former about 3-6 dB down from the latter.

to that from a single driven dipole and reflector of equal length spaced ¼ wavelength away. The conjecture that this antenna acts like nine stacked reflected dipoles, one above the other, seems to be

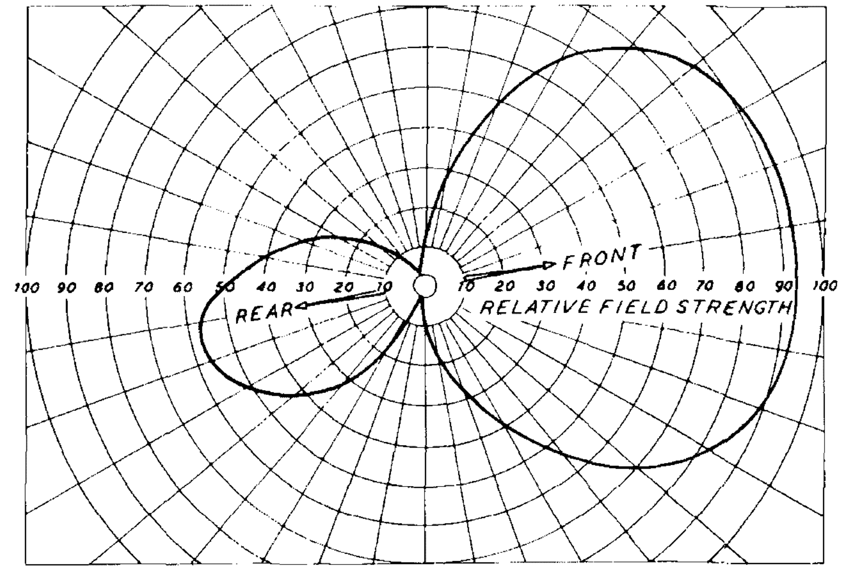


fig. 6. Azimuthal-field pattern of 1/9-scale model 9-element colinear.

The model antenna was mounted by its mast onto a pivoting lazy susan (temporarily sacrificed for the cause) on which a cardboard setting circle was glued. A 1N21 diode was connected in series with the feed point. Two rf chokes (6 turns of hookup wire, ¼ inch dia.) were then connected in series with 300-ohm twin lead, which fed the rectified rf to the foot of the mast where a sensitive current meter was placed near the setting-circle base.

The APX-6 transmitter was coupled to a 4-foot parabolic dish, (F/number 0.25) with a vertically polarized dipole feed. The dish illuminated the scale-model colinear, which acted as a receiving antenna. Readings were taken as a function of azimuth angle of the model.

conclusions

This antenna provides about 8-10 dB gain over an isotropic radiator in the horizontal direction. The metal mast (as borne out by experiment) apparently acts as individual reflectors for each of the nine dipoles. This yields an azimuthal pattern and 3 dB azimuthal gain, similar

confirmed at least in terms of its azimuthal patterns as seen in fig. 6. The analogous situation in terms of polar patterns is open to question at this time. I believe, however, that the basic polar pattern does not change radically from that in free space, as depicted in fig. 4D.

The antenna was originally built as a five-dipole colinear array and was soon after lengthened to nine dipoles. Good reports have been received to distances of 100-plus miles.

I would like to thank K6OQK and WA6MEM for assistance in measuring the azimuthal patterns of the model. W6MMU loaned me his APX-6 and parabolic dish, which resulted in final pattern measurements of the antenna model.

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ham radio

gamma-loop-fed vertical dipole

Proven on 2-meter fm and the citizens band, the gamma-loop-fed vertical half-wave antenna offers easy erection and tuneup

William I. Orr, W6SAI

Owners of four-element quads and many prominent DXers are quick to denounce the simple ground plane as an antenna "that radiates equally poorly in all directions." Notwithstanding this opinion, many antenna manufacturers are quick to point out their more popular and fastest moving item in the amateur line is the simple, easily erected, low-cost ground plane. In addition, many of the robust-sounding signals from overseas DX stations are radiated from ground planes, according to the information on the QSL card.

That's what makes a horse race. The ground plane cannot be brushed aside as a worthless antenna because many amateurs do use it, and work lots of DX with it. Commercial and other services make use of the ground plane and improved versions of it for vhf point-to-point and mobile services. Many amateurs approve of it as a compact, low-profile antenna, well suited for portable work and for a quick installation that does not irritate neighbors.

the vertical dipole antenna

An improved variation of the ground

plane has been in military and commercial service for some time. It affords somewhat better operation than the common version and does away with the usual radial system. This version consists of a half-wavelength vertical dipole radiator with a "ground independent" feed system.[1] Various versions of this antenna system are shown in **fig. 1.** It is well suited to either fixed or mobile operation.

The half-wavelength dipole provides about a 2-dB gain over the usual quarter-wavelength monopole antenna which is a worthwhile increase in performance for a minor alteration in size.[2] In addition, the usual ground plane radials are not required with a half-wavelength antenna as the isolation from the feedline with this antenna is nearly complete if a perfectly-balanced center-feed system is used. The vertical dipole antenna tends to run into trouble, however, when end-feed is used, and the feed system is unavoidably

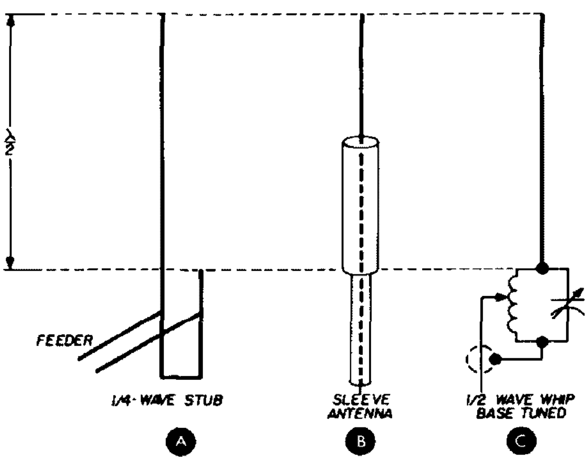


fig. 1. Simple end-fed vertical dipole antennas. (A) Quarter-wavelength stub and open-wire feedline were used in early thirties. Radiation from stub and line interfered with normally low-angle radiation of antenna. (B) Improved sleeve antenna used lower portion of dipole to act as shield for coaxial transmission line, reducing effect of line coupling to a great extent. (C) Vertical dipole matched to a low impedance coaxial line by means of a tapped resonant circuit.

coupled to the field of the antenna; antenna currents on the feedline tend to make it a part of the radiating system, substantially altering the antenna field pattern.

The free-space radiation pattern of the ground plane and the vertical dipole are

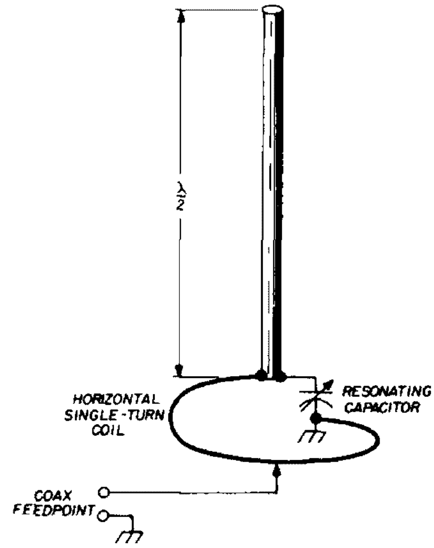
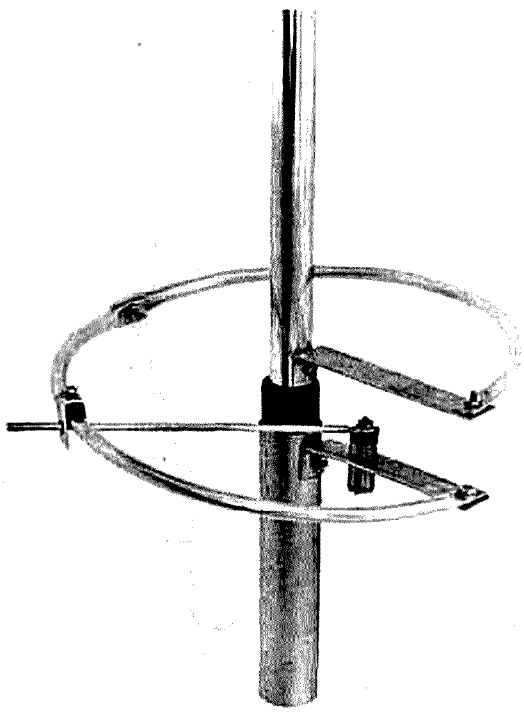


fig. 2. The gamma-loop-fed vertical dipole. Parallel-resonant circuit is mounted in the horizontal plane to reduce coupling to the antenna to a minimum value. Single-turn coil is tapped at appropriate spot for feedline. A practical version of this antenna is shown in the photograph.

equivalent figure-8 plots in the X-Y (vertical) plane provided that the feed system and antenna are not coupled to each other.[3] The pattern of either antenna close to the ground, then, is a function of ground reflection. Typical reflection patterns for these antennas are shown in the *ARRL Antenna Book*, and it can be observed that a reasonably low angle of radiation is obtained when the bottom of the vertical dipole antenna is less than a half-wavelength above ground.[4] This gives the user plenty of latitude in mounting the antenna, especially in view of the fact that the theoretical ground reflection patterns are distorted by the real-life environment surrounding the antenna.

The free-space radiation plots show that a vertical half-wave antenna may be mounted with its center as much as three-quarter wavelength above ground before the vertical radiation pattern begins to rise appreciably. For the 20-meter band, this means that the bottom

of the vertical may be ½-wavelength, or about 35 feet above the ground surface. The vertical could therefore be mounted atop the roof of a one- or two-story



Cush-Craft gamma-loop-fed vertical dipole for 10 and 11 meters has loop broken into two parts for ease in shipment, and capacitor is made up of dielectric sleeve between two sections of vertical tubing. Gamma rod reaches from coaxial plug at right to moveable clamp on inductor at left.

house without apprehension. In view of the character of the ground around typical amateur installations, moreover, and considering the influence of nearby objects, the free-space radiation plot can be considerably blurred, and the half-wave vertical has given good DX results when mounted at heights of over 80 feet above ground (atop a 4-story building, for example).

Conventional wisdom has it that the radials on a ground plane lower the angle of radiation of the antenna, and that the more radials employed, the lower the radiation angle. A more correct view is that the radials decouple the outside of the coaxial transmission line from the *ground plane antenna* and provide a low impedance ground point for the outer shield of the transmission line at the base of the quarter-wave antenna. If the radials are missing, the line acts to provide the missing quarter-wave section of the antenna, or more, if the line has appreciable length. The antenna plus transmission line, then, is transformed into a long vertical antenna having a high angle of radiation, most of which is useless for long distance communication.

a practical feed system

The feed system for the half-wave vertical antenna must provide a good impedance match between the high-impedance antenna end-point and the low-impedance transmission line, and must decouple the outside of the line from the field of the antenna. These two requirements can be met without the use of radials by the technique shown in fig. 2. A parallel-tuned, gamma-matched resonant circuit is used, positioned in such a way as to offer minimum coupling to the antenna element. The inductor is a large, horizontally mounted, single-turn loop and the parallel capacitor is a fixed, high-voltage ceramic unit. By varying the length of the antenna, the size of the coil and the tap point on the coil, a good impedance match may be achieved for either a 50- or 70-ohm coaxial transmission line.*

To insure that the coaxial line is not coupled to the antenna once it leaves the tuned circuit, the line should ideally be led away at right angles to the antenna. This is usually not practical, so a ferrite rf choke is placed along the transmission line about one-eighth wavelength away from the coupling circuit. The choke is simply made up of two *turns of the* transmission line threaded through a large ferrite core. Induced currents on the outer shield of the line attempting to turn the line into a long-wire antenna are

*The antenna shown in the photograph is manufactured by Cush Craft, 621 Hayward Street, Manchester, New Hampshire 03103, and is intended for either 10- or 11-meter operation.

suppressed at this point (**fig. 3**).

The single-turn inductor and capacitor are resonant at the operating frequency of the antenna. For the antenna illustrated, this is 28.6 MHz. For those wishing to roll their own, a nomograph

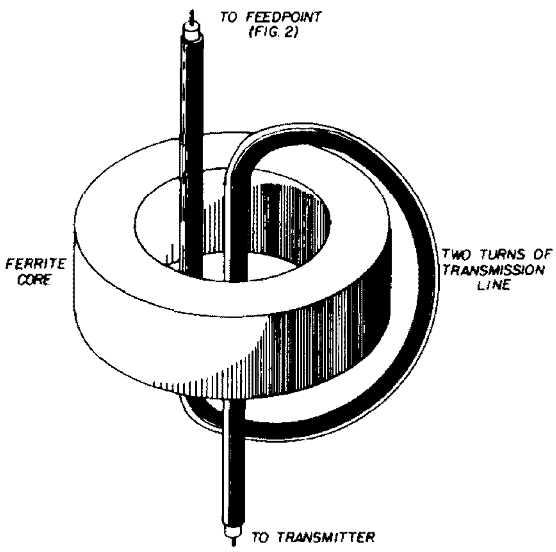


fig. 3. Coaxial line may be decoupled from field of antenna by looping the line through large ferrite core, making simple rf choke in series with outer shield of line. Choke is placed a few feet below the antenna. A suitable core is the Indiana General CF-123 toroid, 2.4 inches in diameter (Q-1 material having a permeability of 125 at 1 MHz). Available as catalog number 59F-1518 from Newark Electronics Corp., 500 North Pulaski Road, Chicago, Illinois 60624. Transmission line is looped through core twice, but is shown as only once, for simplicity of illustration.

for determining the inductance of a large single-turn inductor is given in **fig. 4**.[5] The inductance value should be chosen so as to resonate the tuned circuit to the operating frequency with a capacitor whose value in picofarads is about two to four times the wavelength in meters.

The ten-meter antenna in use has a 50 pF ceramic capacitor, with an inductance of about 0.7 μH. The coil is made of 3/8-inch aluminum tubing, about 11 inches in diameter. The coil and capacitor combination are grid-dipped to frequency on the bench before being mounted on the antenna.

antenna adjustment

The dipole is cut to one-half wavelength by formula and connected to the pre-tuned base network. The antenna may be mounted in the *clear*, about head-height on a step ladder for initial adjustment. The tap on the base inductor is moved back and forth, a half-inch at a time until the point of lowest swr is found. This is done while feeding a few watts to the antenna from the station exciter. If this adjustment does not produce a low value of swr (say, less than 1.5 to 1), the length of the whip or the diameter of the coil may be varied a bit to bring the whole system into resonance. The tap, antenna length and coil diameter are all interdependent. But with the information given, you are in the ball park and only minor tweaking is needed to bring the swr to near-unity at your design frequency. Once the correct adjustments are determined, they are locked, and the antenna is ready to place in the final operating position.

no radials

Properly isolated from ground, the vertical half-wave dipole requires no radials. The use of the line choke and positioning of the feedline beneath the antenna insures minimum coupling between line and radiating element. A quick test of line isolation may be made by clipping a single quarter-wave horizontal radial to the coaxial shield of the line at the base of the antenna. If adding the radial changes the swr to an appreciable extent, parasitic coupling exists between the line and the antenna and the addition of one or two radials is suggested. If the addition of the temporary radial does not alter the swr, it is not needed. Leave it off.

operation

The half-wave vertical antenna provides about 2 dB gain over the groundplane antenna and places the point of maximum current a worthwhile distance higher in the air, assuming the same base mounting point as a conventional ground

plane. In addition, radials are not required to make it work properly. This is all to the good. Such an antenna has been used with some success at the author's portable QTH in Hawaii (KH6ADR) and,

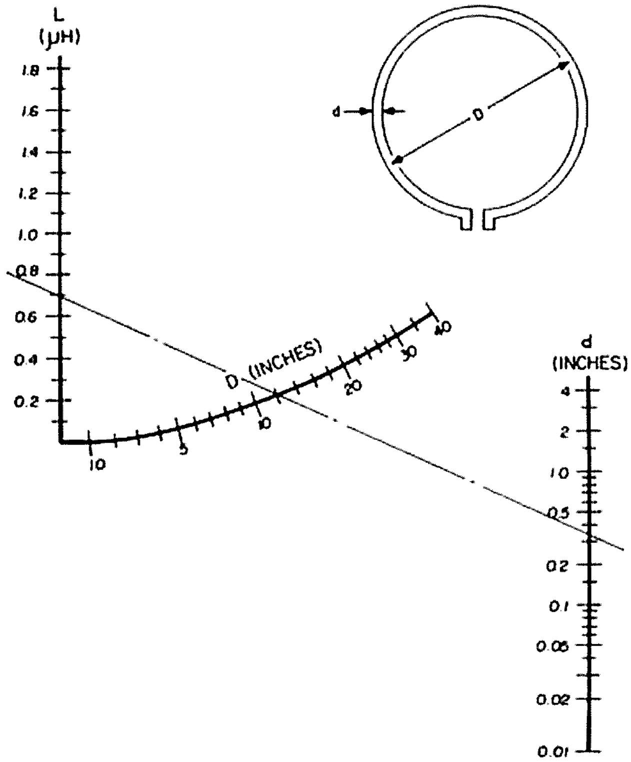


fig. 4. Circular loop inductance nomograph. An 11-inch coil of 3/8" aluminum tubing, for example, exhibits about 0.7 μH inductance.

while comparisons are really invalid, it seems to out-perform the ground plane previously used. In any event, DX can be worked and the antenna functions well and is very inconspicuous. What could be a better testimonial than this for such a simple and unpretentious installation?

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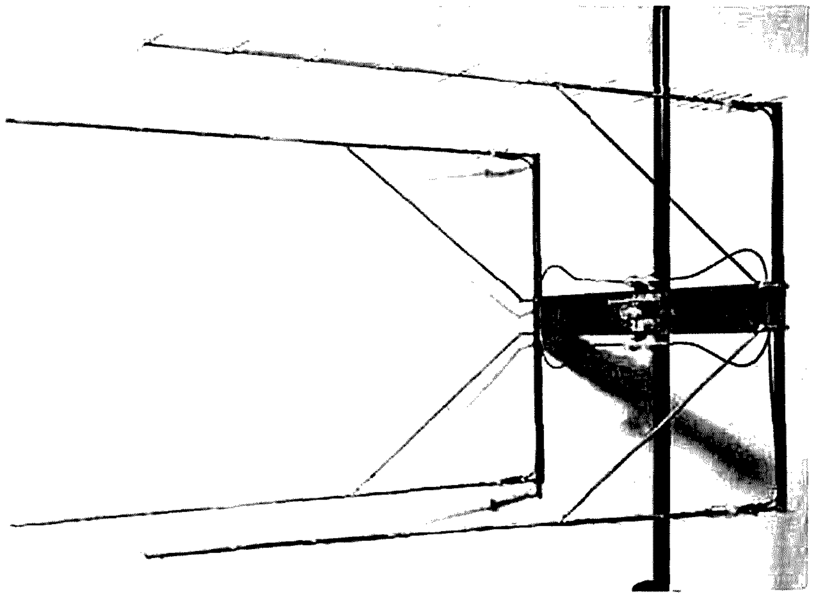
ham radio



a successful 1296-MHz Yagi

Here is a 1296-MHz quad Yagi which overcomes the usual construction problems and provides gain equivalent to a four-foot dish

Reed E. Fisher, W2CQH, 2 Forum Court, Morris Plains, New Jersey

It is well known that the Yagi is probably the best all-around antenna for vhf. It is easy to build, provides high gain with low weight, and offers low wind resistance. However, it is often hard to scale down a Yagi to work on the 432- or 1296-MHz bands. The poor performance often experienced on these frequencies has led to the erroneous assumption that "Yagis won't work at uhf."

I feel that the much-publicized poor performance of the uhf Yagi comes about because the dimensions given for director lengths—the most critical part of a Yagi—simply do not scale down properly. For example, if the experimenter simply scales a two-meter Yagi down to 1296 MHz, without doing any additional measurements, he will usually obtain mediocre results.

design

The 1296 MHz, 13-element antenna shown in fig. 1 is a scaled version of Orr and Johnson's famous long Yagi, which, for the past eight years, has given outstanding results on 144 and 432 MHz at my station.[1] The parasitic elements are

lengths of AWG 14 copperweld wire soft-soldered to a boom made of ¼-inch OD brass tubing. The copperweld provides both physical strength and high electrical conductivity, the second condition being required for effective Yagi operation. The director lengths, the most critical dimension of the antenna, have been arrived at by careful pattern measurements on an antenna test range.

A few words of caution: Do not use directors or boom of different diameter than specified. Also, do not use an insulated boom. Any of these modifications will detune the directors.

driven element

The driven element, shown in fig. 1 and the photo is a delta-matched dipole also constructed of AWG 14 copperweld pushed through the boom center and secured by a Plexiglas block and set screw. The delta-matching section consists of two pieces of AWG 14 soft copper wire soldered to the dipole.

The driven element is fed by a half-wavelength loop balun constructed of 0.085-inch OD solid coax. The balun is excited by the main feedline made of 0.141-inch OD solid coax. Both balun and main feedline are soft-soldered to the brass boom. The solid coax is probably the main contributor to the success of this Yagi, since it provides a compact and low loss feed arrangement. The Teflon-dielectric 0.141-inch coax exhibits a loss-per-foot comparable to RG-8/U.

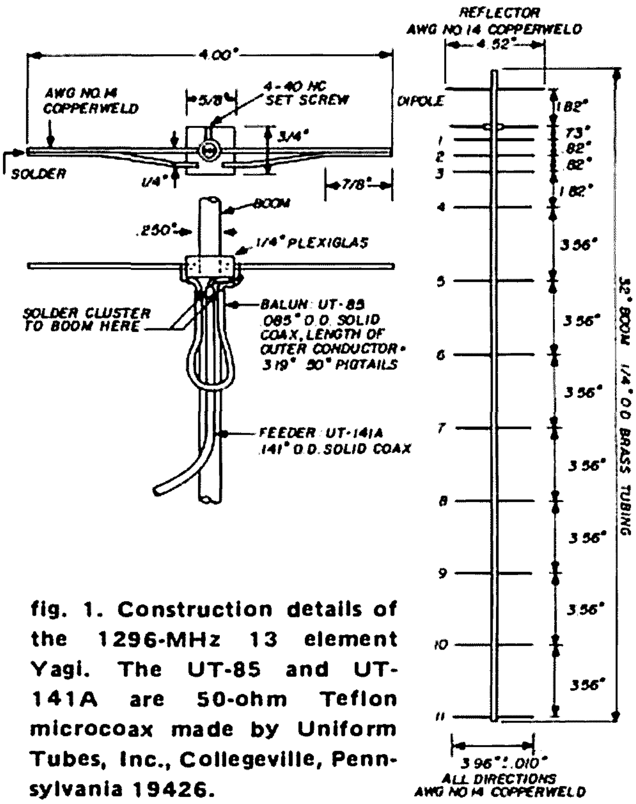
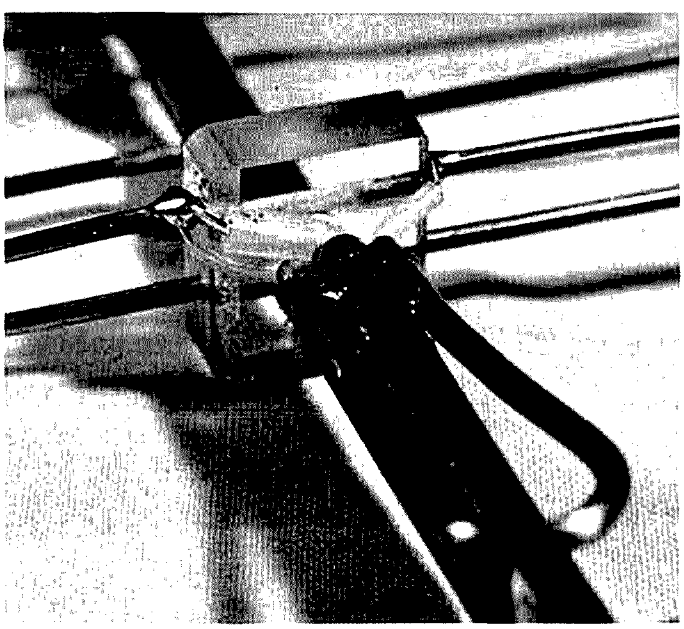


fig. 1. Construction details of the 1296-MHz 13 element Yagi. The UT-85 and UT-141A are 50-ohm Teflon microcoax made by Uniform Tubes, Inc., Collegeville, Pennsylvania 19426.



Detail of the dipole and feed arrangement.

performance

A single Yagi exhibits 15.2 dB gain (over isotropic radiator) when compared with a 15.0-dB gain reference horn antenna. Figs. 2A and 2B show the azimuth (E plane) and elevation (H plane) patterns of the Yagi, measured at an industrial laboratory test range by R. H. Turrin, W2IMU. Note that in fig. 2A the first sidelobes are down about 9.5 dB, a condition necessary for effective Yagi operation. If the builder has any doubts about the correctness of director lengths, he should measure the first sidelobe levels. If they are not down by at least 9 dB, the Yagi is not working properly.

When the patterns of figs. 2A and 2B were integrated in 1° increments, antenna directivity was found to be 17.55 dB. This number corresponds closely to that given by the well known approximation,

$$\text{directivity} = 10 \log \frac{41253}{\theta_E \theta_H} = 17.70 \text{ dB}$$

where θ_E and θ_H are respectively, the -3 dB E-plane and H-plane beamwidths in degrees. Since antenna directivity and gain differ by the element skin resistance losses, we must conclude that for this Yagi these losses amount to about 2.5 dB. This emphasizes the importance of building the antenna elements from low-loss metals such as copper.

The single Yagi should exhibit a vswr not exceeding 1.2:1 (with respect to 50 ohms). The vswr will increase to above 2:1 when droplets of water cling to the dipole, but will recover to 1.2:1 when the antenna is shaken or allowed to dry.

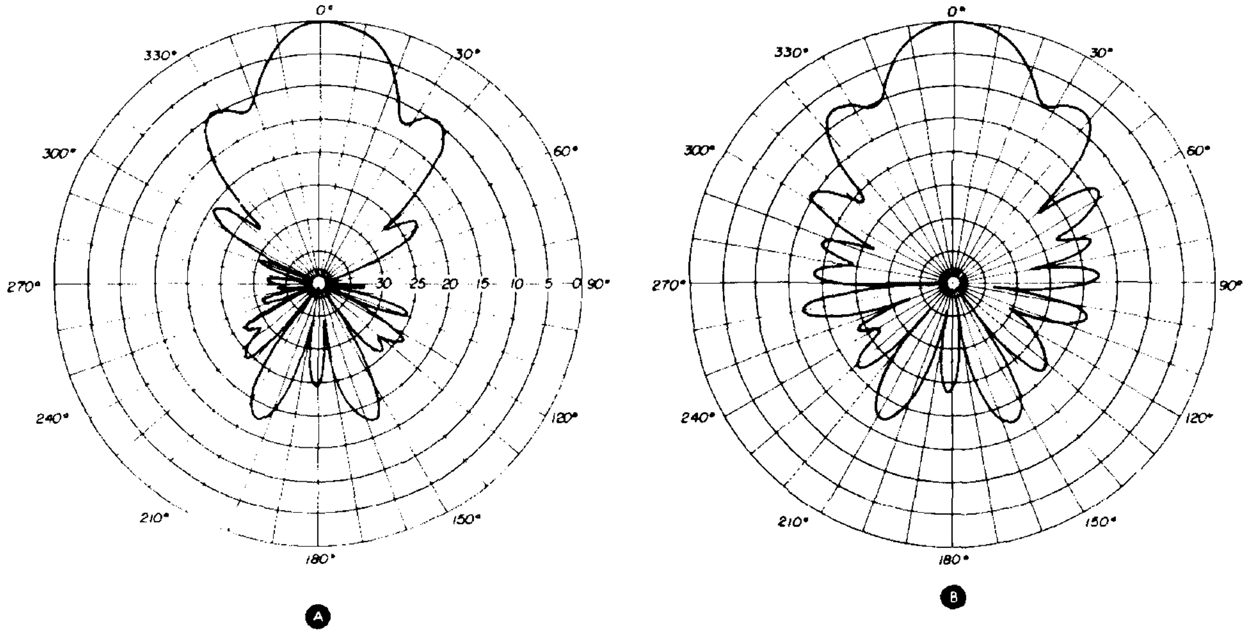


fig. 2. Azimuth pattern of the antenna.

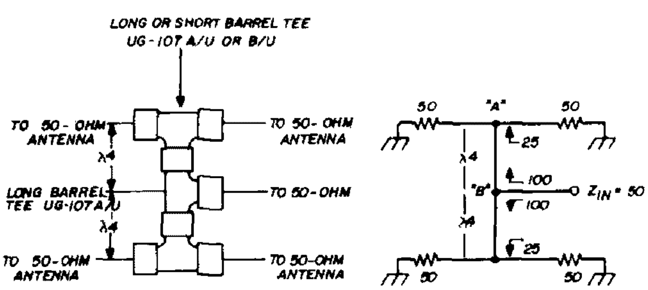


fig. 3. Matching network is made up of three, long barrel UG-107A/U coaxial tees. The matching schematic is also shown.

matching

My final quad Yagi array uses 20-inch spacing between antennas. This array provides 21-dB gain which is comparable to a four-foot dish. The matching network, located at the array center, is simplicity itself and is shown in **fig. 3** Three type-N coaxial tees joined together will match four 50-ohm antennas into one 50-ohm generator with low vswr at 1296 MHz.

The matching theory is shown in the schematic diagram. The two 50-ohm loads (antennas) at the top of the diagram appear in parallel at *point A* and, thus, look like 25 ohms. This 25 ohms is transformed up to 100 ohms at *point B* by the quarter-wave-length 50-ohm matching section. The two 100-ohm loads at the *B junction* combine in parallel to again form 50 ohms, hence, the array is matched.

reference

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ham radio

direct reading and expanded scale swr meters

An inexpensive surplus meter presents a novel method to read forward power, reflected power, and swr — simultaneously

Earnest A. Franke, WA4WDK, 108 Matawan Terrace, Matawan, New Jersey 07747

Most amateur radio stations use an swr meter to indicate the approximate standing-wave ratio of their antennas. The meter also indicates the relative output power of the transmitter. The standing-wave ratio is the ratio of the reflected power to the forward power.

A signal proportional to each of these powers comes from a directional (parallel wire) coupler with rectifier diodes. Most swr meters must be adjusted in sensitivity so that the voltage corresponding to the forward power will give a full-scale meter deflection. The meter is then switched to monitor the reflected signal, which will be less than full scale. The ratio of the reverse meter reading to the forward meter reading yields the proportional meter reading, M.

$$M = \frac{\text{reverse meter reading}}{\text{forward meter reading}}$$

The standing wave ratio may then be computed, based on a linear detector.

$$\text{swr} = \frac{1 + M}{1 - M}$$

The limitations, such as return feed-line loss, diode nonlinearity, and coupler directivity have been pointed out in past articles.[1, 2, 3] Because of its accuracy, the swr meter is used primarily to indicate relative improvements in matching. The meter described here will be as accurate as the above limitations permit. The meter does not need to be normalized. Also, a scale-expanding technique will be described.

single meter indicator

Many amateurs have monitored forward and reflected power by using two meters. You may adjust a matching network to decrease the reflected power while constantly watching the forward reading. The obvious extension of this approach is to combine the two meter

movements into one case. The relative power scales remain while the needle crossing indicates the ratio of the values.

I modified an I-101 glide path indicator meter to display these measurements. Glide path receivers are used to guide pilots while making their final landing approach in instrument flight weather. Several of these cross-point indicators are available in surplus for less than five dollars.* Any stops restricting needle movement from twenty degrees on either side of the rows of luminescent dots must be removed. The small counter-balancing needle weights must be bent inward slightly. Any manipulation of the bearing or meter movement must be avoided.

A scale suitable for pasting on the meter face is included in **fig. 1.** The zero adjust is varied so that each needle corresponds with the zero-power indications. The resultant indicator resembles the Bird model 3122 Thruline wattmeter. An swr of 1:1 is obtained if the reflected power is equal to zero. Conversely, an infinite swr exists if the reflected power is equal to the incident power.

The scale of **fig. 1** *is based on the* 1.31 power law detector response, discussed by Jerry Hall, which approximates most

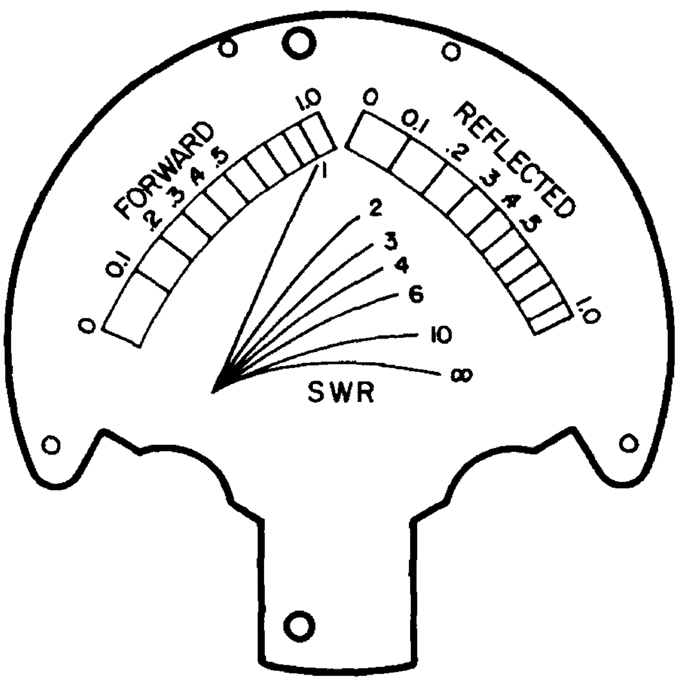

fig. 1. An swr scale to use on the cross-point indicator without the meter expander.

detectors.[2] A more exact standing-wave ratio formula would then be

$$\text{swr} = \frac{1 + \sqrt[1.3]{M}}{1 - \sqrt[1.3]{M}}$$

The various ratios for normalized meter

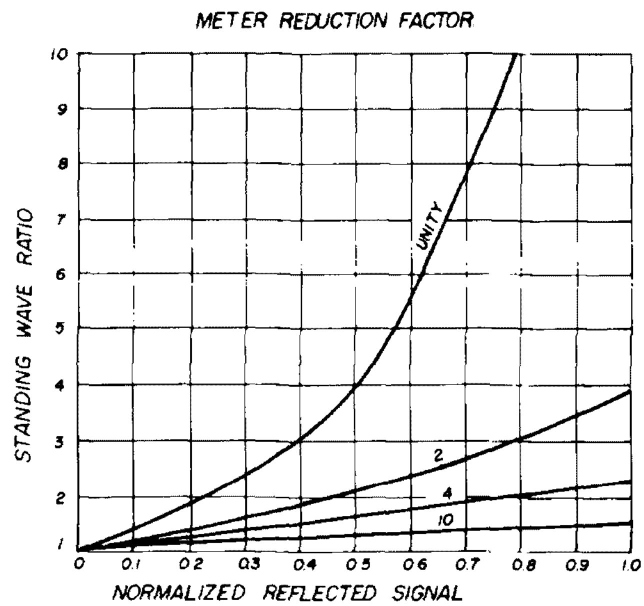

fig. 2. Graph shows the expanding effect of reducing the forward signal.

readings are shown in **fig. 2.** The curve labeled unity is the one to use without meter-expanding devices. The schematic of the cross-point indicator with the optional expander is given in **fig. 3.**

scale expander

Most amateurs must turn down the sensitivity of their swr meters, especially at increasing frequencies because the length of the sampling *wire becomes* a greater portion of the wavelength as the operating frequency is raised. If the signal corresponding to the forward power were attenuated, the sensitivity control would have to be increased. When, in a conventional meter, the meter is switched, the reflected signal will deflect the needle further.

Swr-coupler operation is based on the

*Fair Radio Sales Company, Box 1105, Lima, Ohio 45802 sells government reconditioned I-101 cross point indicators for $2.95 plus shipping.

fact that the sampling diodes are nearly matched to each other. Each diode drives the same or equal resistances. The response of a detector is affected by the meter resistance. If part of the forward-power meter current is shunted by a resistor, the resistance seen by the diode decreases. This shunting effect must be compensated by a series resistance in order to match the load through which the reverse signal current will flow. If a resistance equal to the meter resistance is placed in parallel with the indicator meter, the needle will swing to one-half its former value. A meter reduction factor, N, of two is achieved. A series resistor of one-half the meter resistance must be added to the shunt combination

$$R_{SH} = \frac{R_M}{N - 1}$$

where R_{SH} = shunting resistance

R_M = meter resistance

N = meter reduction factor

$$R_{SE} = \frac{(R_M)^2}{R_M + R_{SH}}$$

The value of M, the normalized reflected signal, is increased by the meter

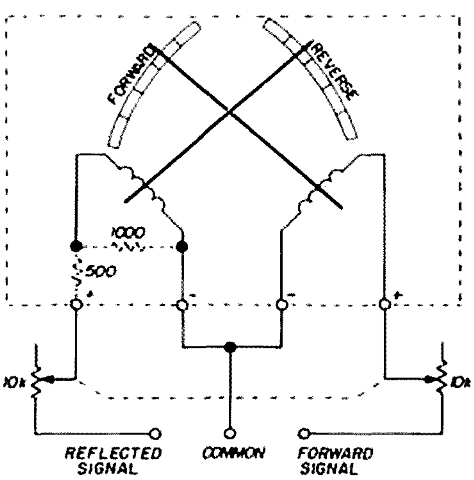

fig. 3. Cross-point swr indicator that eliminates the normalizing or calibrating action. Two ganged 10k pots provide sensitivity control.

reduction factor.* A new normalized reflected signal ratio is indicated when the reflected signal is monitored.

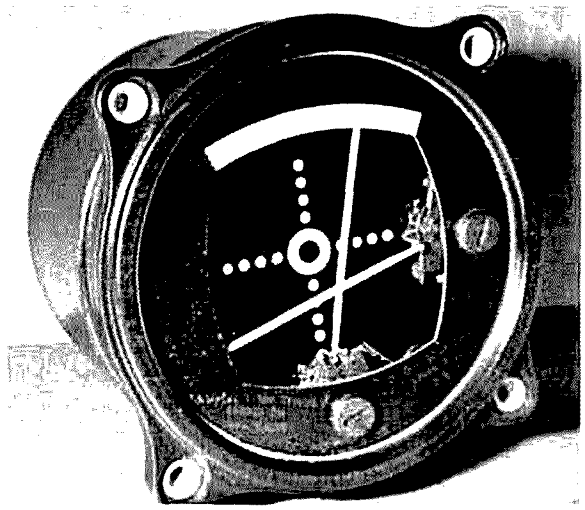
Photo of the Signal Corps I-101 cross-point indicator.

$$M = \frac{(\text{M indicated})}{(\text{meter reduction factor})}$$

This new value may be inserted in the previous swr formula.

An example is shown that might be applied to most swr indicators. Suppose the meter has been adjusted to full scale while monitoring the forward signal, and the reverse signal is to be measured. A half-scale deflection, M = 0.5, corresponds to an swr of 4:1. If a resistance equal to the meter resistance is added in shunt, and another resistor added in series, the meter will be deflected down to one-half the previous reading. The sensitivity control must be turned up so that the forward signal will still deflect full scale. The reflected reading will also double. The full-scale reading has thus expanded to a 4:1 swr indication of full scale. This assumes that there is enough signal to provide a full-scale deflection in

*In normalized readings, one key value is called *1*, and all other readings are related to this. In using Smith charts, for example, 50 ohms is the reference and is called *1*. 70 ohms becomes *1.4* and 30 ohms becomes *0.6*. If, in swr measurements, you call a forward reading of 25 watts *1*, a reflected reading of 5 watts becomes *0.2*. If the actual reflected reading was written as R = 5, the normalized reading would be indicated by R' = 0.2 with the prime signifying a normalized reading. For minds conditioned to the decimal system and percentages, normalized readings are often a handy tool. editor.

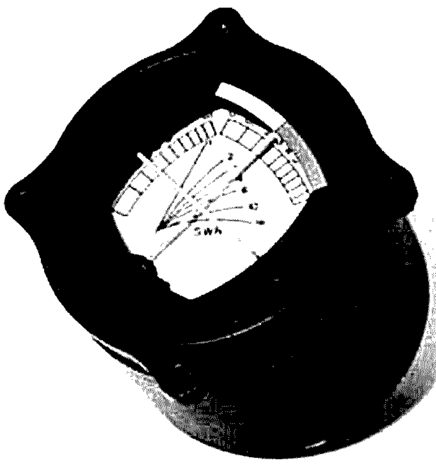

Photo of the I-101 converted to a direct-reading swr meter.

the forward position. Fig. 2 shows the expanding effect of reducing the forward signal.

The correlation between swr and the reflected signal reading becomes linear as the forward signal is reduced. If you have a large signal available from the swr coupler and a low standing wave ratio, the expander will allow a magnified reading of the meter. Fig. 4 shows a

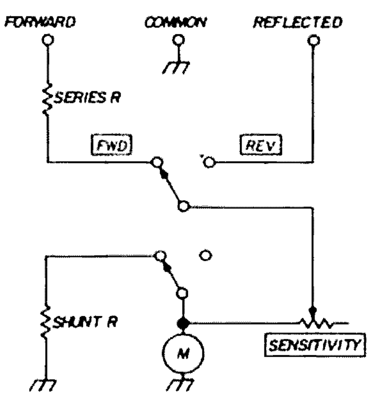

fig. 4. Modification to a conventional swr indicator to expand the scale.

schematic of the expander added to a conventional swr indicator.

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ham radio

an all-band phased-vertical antenna system

Phased groundplane verticals offer directivity, multiband operation and mounting flexibility without buried radials

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Endeavoring to design an antenna system for my home, I had a stiff list of requirements. I wanted:

1. Operation on 80 through 10 meters with reasonable efficiency and low swr at all operating frequencies,

2. Small physical size to fit on a postage stamp size lot (not large enough to support a long wire, dipole or inverted-V),

3. Selectable antenna directivity to enhance medium-power operation (200 watts) and reduce interference,

4. Low cost—precluding towers and rotators, and

5. Reasonable esthetics in deference to good neighbors.

No available antenna system is known to meet all of these requirements, hence the design of this system has involved a series of compromises. It is felt, however, that the system to be described is a reasonable compromise applicable to other locations. This system has been in use at my station since the spring of 1970, and its operation has been proven in many contacts on all bands, including some DX.

theory and design

Since dipoles, inverted-Vs, long wires and beam antennas were not consistent with the system constraints either because of physical size, cost or non-selectable directivity, the first approach was to examine vertical antennas. A vertical antenna does not require a lot of space, does not cost much money and is reasonably esthetic. In addition, recent articles have discussed the advantages of operating two verticals at various phase angles to obtain varying antenna patterns.[1,2,3] Generally speaking, feeding both antennas in phase (0° phase angle) results in an antenna pattern favoring a line at right angles to the plane of the two antennas (broadside pattern), whereas feeding one antenna out of phase (180° phase angle) results in a pattern favoring a line in the plane of the two antennas (endfire

pattern), and feeding one antenna 90° out of phase favors a line off one end of the plane of the two antennas. Other phase angels produce other patterns generally less useful for amateur purposes. In these previously published articles, the phase angle is achieved by feeding each antenna with a different length of coaxial cable.

For my specific application, two phase angles are of interest: 0° and 180°. These two phase angles produce antenna patterns which are rotated 90° from each other: north-south and east-west. Feeding both antennas with the same length feedline produces the 0° phase angle and its broadside pattern, and is suitable for multi-band operation.

Two methods can be used to generate the 180° phase shift needed for endfire operation: make the feedline to one antenna ½-wavelength longer than the feedline to the other antenna, or reverse the conductors of the feedline to one antenna. The first method is generally used for single-band operation, but is not feasible for multi-band use because there is not a ½-wavelength difference between the two feedlines on the different bands, and, hence, not always a 180° phase shift. For example, a ½-wavelength difference on 40 meters (180°) is only ¼-wavelength difference (90°) on 80 meters. Although not employed here, by properly selecting antenna spacing and feedline dimensions the 180° phase shift can be maintained for operating frequencies which are odd multiples of each other—such as 40 and 15 meters.

The second method of obtaining the 180° phase shift is not possible for coaxial fed antennas because, in a normal installation, the feedline is unbalanced with one side of the feedline grounded. This method is feasible, however, if the antennas are fed with a balanced transmission line like open-wire feeder. I chose the system of feeding two ground plane vertical antennas with open-wire feeders of equal length and having the ability to remotely reverse the feedline to one antenna. Open-wire feeders necessitated an antenna tuner, but allowed the construction of vertical antennas without the usual coils or traps. Appropriate coils are required for the antenna tuner, however. Fig. 1 summarizes my system.

construction

Construction of this antenna system is fairly simple and uses readily available materials. The only critical factor is retaining a fair degree of symmetry in the construction and installation of the two antennas so that *predictable* antenna patterns will result and each antenna will carry half of the radiated power. The construction details given are illustrative only; they can surely be improved. Since my system was designed to fit on my house roof, the physical size and mounting details were developed accordingly. The size of each ground plane antenna was primarily dictated by the eve-to-eve dimensions of the roof. I used 25 feet for the vertical element and for the ground plane radials. The base of each antenna was supported ten feet above the peak of the roof.

I used thin-wall electric conduit for the vertical element, and number 16

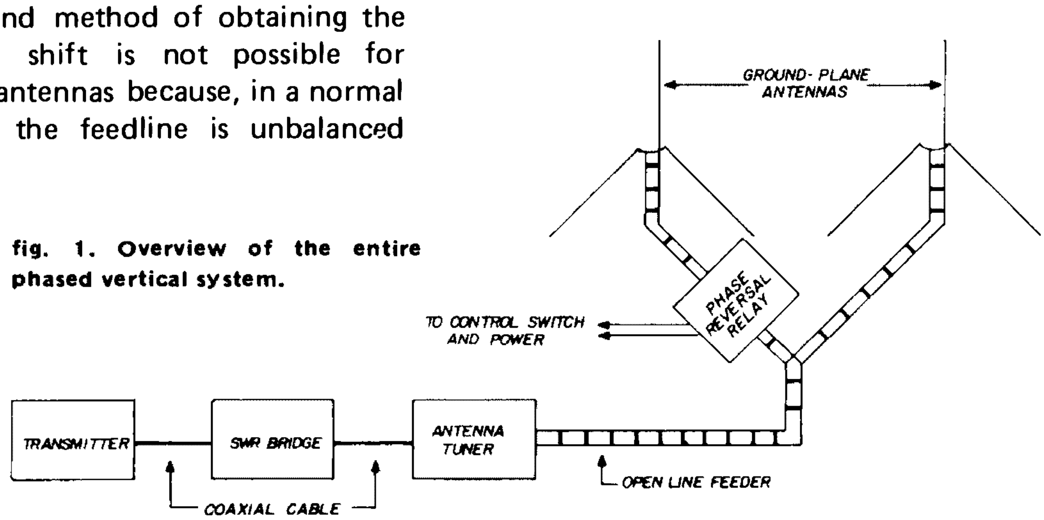


fig. 1. Overview of the entire phased vertical system.

table 1. Theoretical system gain with an antenna spacing of 35 feet.

frequency (kHz)	wavelength spacing	antenna broadside	gain dB endfire
3,750	0.15	0.5	4.2
7,150	0.28	1.2	3.7
14,200	0.55	4.5	2.0
21,250	0.80	4.5	2.0
29,000	1.10	2.5	4.0

antenna wire for the radials. Rigidity dictated ¾-inch conduit. Since conduit is usually sold in ten-foot lengths, the vertical element must be spliced to achieve the 25-foot overall length. The splices must be mechanically rigid and electrically conductive. I used five-foot lengths of ½-inch conduit inside the ¾-inch conduit at each joint. Some shimming is necessary to achieve a rigid splice.

A ten-foot length of ¾-inch conduit was also used to support the antenna off the peak of the roof. The ground plane radials also acted as guy wires at the ten-foot level. Non-conductive guys (I used 50-pound test fishing line) are used at the thirty-foot level. **Fig. 2** details the construction and mounting of the antennas.

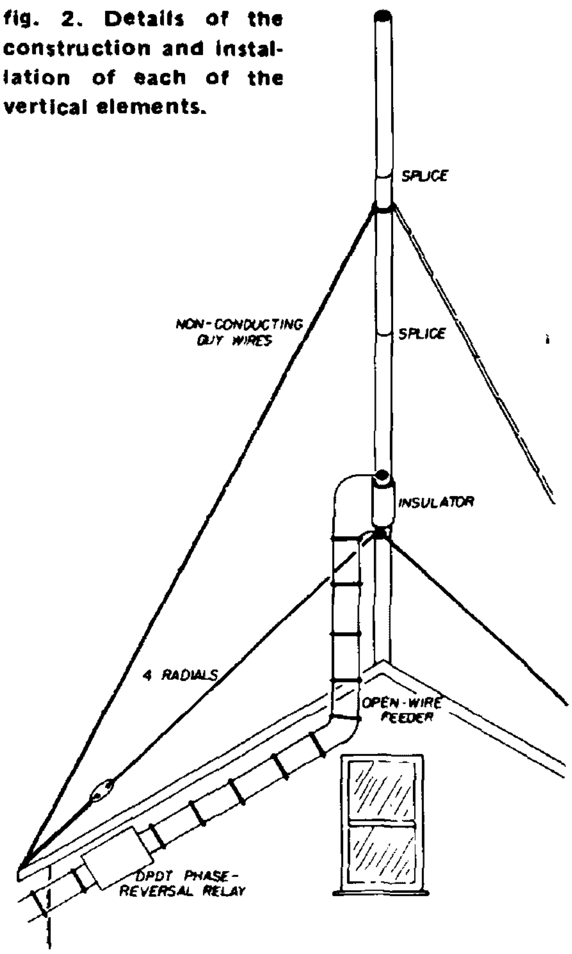
phase-reversal system

Insert a double-pole, double-throw relay in the feedline to one antenna for the phase reversal and antenna pattern shift. I used a surplus relay with ceramic insulation and ¼-inch contacts, but any relay with large enough contacts and insulation to handle the transmitted power should be adequate. Be careful when installing the relay so that the feedline length is not significantly altered as the relay is operated. In addition, the feedline spacing should be maintained through the relay. Enclose the relay in a plastic box for protection from the weather. The relay control line should be run at 90° from the feedline for as long a distance as possible.

antenna tuner

The basic tuner design was taken from the *ARRL Handbook,* and utilized series and parallel tuning to match a wide range of antenna reactances.[4] Two band-changing options are available: plug-in coils for each band, or one set of coils with appropriate taps. I chose the latter option because the L/C ratios for matching the antenna on each band could be determined by experimentation. Either a heavy duty rotary switch or alligator clips can be used to change bands. **Fig. 3** is a schematic of the antenna tuner. Note that C2 and C3 must be isolated from ground and from each other. In constructing this tuner, I used the variable capacitors from

fig. 2. Details of the construction and installation of each of the vertical elements.



an ARC-5 transmitter for C2 and C3, removed the worm gear drive screw, and drove both capacitors in tandem with an insulated fibre gear.

orientation and spacing

Antenna gain for both endfire and broadside arrays is a function of the wavelength spacing between the two driven elements.[5] If the spacing is either

too close or too far apart, the antenna has either no directivity and no gain, or it becomes very directive with a very narrow pattern. Since this system was designed for general amateur operation, fairly broad patterns (90°) with moderate gain on all hf bands was desired. Available space may dictate the spacing, however. I used 35 foot spacing at my station. Table 1 summarizes the *theoretical* system gain for each hf band (on-the-air contacts have verified the relative gains between broadside and endfire patterns).

Since this antenna array is directive, and since its gain and pattern width varies for the different bands, the antenna system spacing and orientation should be chosen by determining the direction of the desired patterns for each band.[1] At this location, for example, the 80- and 40-meter performance is adequate for all stateside contacts, and so the antenna was oriented to optimize directivity towards Europe (endfire) and Asia (broadside) on 20 meters. The orientation of the antenna is, at best, a compromise.

operation

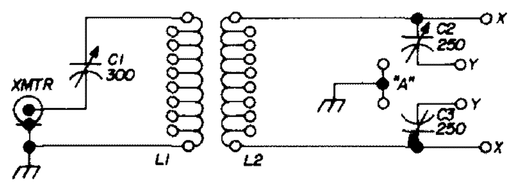
An swr bridge is essential for the proper operation of this antenna system. Proper antenna tuning is indicated by minimum swr and maximum power out. There is considerable interaction between the antenna tuner and the output circuitry of the transmitter, so adjustment of both sets of controls is required. Basic steps in system tune-up are:

1. Select an operating frequency. Select the proper antenna coil taps for resonance (if the taps are properly determined, then one set of taps will cover a complete amateur band).

2. Determine the proper phase angle to enhance radiated power in the desired direction.

3. Adjust the antenna tuner for minimum swr. It is suggested that this step be performed at low power since a very high swr may be encountered initially. Adjust the transmitter output circuitry for maximum power out. Readjust the antenna tuner for minimum swr. It should be noted that additional adjustment of the antenna tuner may be required as the phase angle is changed, even if the operating frequency is not changed.

The antenna system has lived up to its design expectations. It has provided de-



C2,C3	250 pF. Salvaged from ARC-5 transmitter
L1	8 turns, 6 turns per inch, 2-inch diameter. Tapped each turn. L1 is placed inside of L2 with the taps of L1 coming out between the turns of L2
L2	40 turns, 6 turns per inch, 2¼-inch diameter. Tapped 2, 4, 10 and 20 turns each side of the center.

fig. 3. Schematic of the antenna tuner. Connect the antenna at the points marked "X" for parallel tuning; "Y" for series tuning. When using parallel tuning, connect the grounded jumper "A" between C2 and C3.

sired directivity and gain on all hf bands, yet its patterns are broad enough to assure that I can work all directions. The extra chore of adjusting the antenna tuner can be bothersome, but the system's overall advantages more than compensate for this problem. I am sure that this system can find use at other stations.

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ham radio

small-loop antennas

Radiation patterns, design and theory for an effective limited-space antenna

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A small loop antenna is any constant-current loop with a perimeter not exceeding about one-third wavelength. Larger loops can be made to have a constant current around the perimeter through the use of appropriately-spaced phasing devices. The characteristics of such loops, however, are quite different from those of the small loop, and depend, rather critically, on the precise shape of the loop.

Most of the advantages of the small loop are pretty obvious from this definition. Most important of all is that the loop takes up considerably less space than the corresponding half-wave linear antenna. Also, since the characteristics are determined chiefly by the area of the loop, rather than by the shape, they can be made round, square, or any other convenient shape with little change in performance. Among their disadvantages is very narrow bandwidth and disconcertingly low radiation resistance.

Less well known is the effect of the proximity of a large reflecting plane, such as the ground – particularly at the 3.5- to 14-MHz range – the ground is never very far away in terms of wavelength, and it therefore becomes an integral part of the radiating system. This causes the field pattern to deviate markedly from the simple doughnut shape which occurs in free space. It also may have a pronounced effect on the radiation resistance.

Ground reflection characteristics are customarily developed by first assuming the ground to be perfectly conducting, and then substituting an image antenna for the ground plane. The problem is thus reduced to that of an array of two similar antennas, which is not only much simpler than a reflection problem – it also has been solved with sufficient generality to apply without modification to loops. Here we shall concern ourselves chiefly with the results of the calculations, rather than with the formulas themselves. The typical examples shown should give a

good picture of the situation. For those who wish to examine other specific cases, the necessary material is included in an appendix to this article.

Figs. 1 through 4 display the directional characteristics of horizontal loops, while figs. 5 through 9 are devoted to vertical loops. For the latter, the pattern

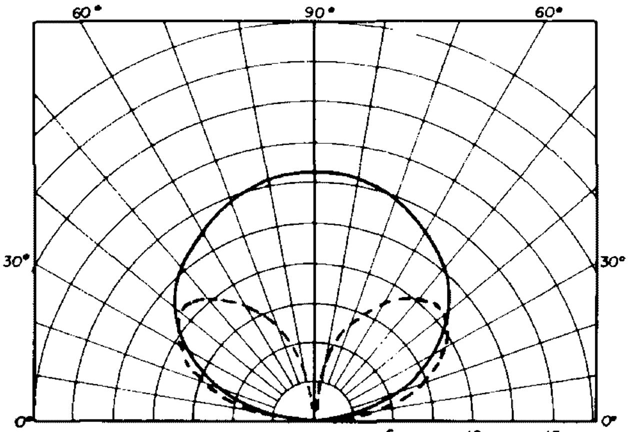

fig. 1. Horizontal loop vs horizontal dipole, both at one-eighth wavelength above ground. For all the patterns in this article, the antenna is assumed to be over a perfectly conducting ground.

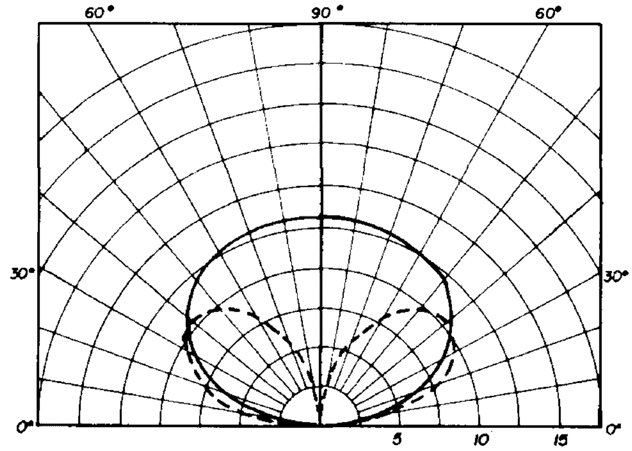

fig. 2. Horizontal loop vs horizontal dipole, both at one-quarter wavelength above ground. In all figures the dipole's pattern is the solid line and the loop's pattern is the broken line.

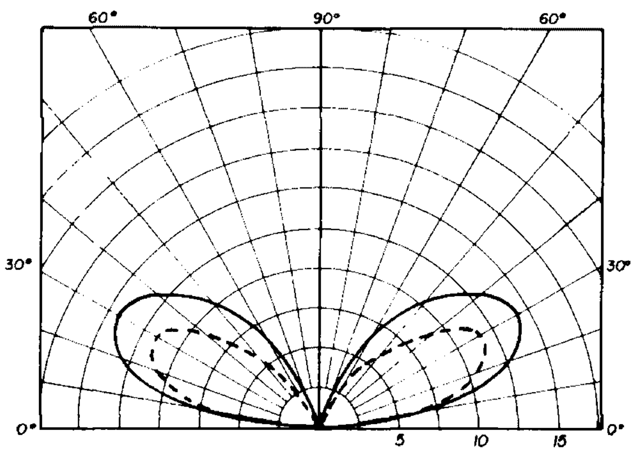

fig. 3. Horizontal loop vs horizontal dipole, both at one-half wavelength above ground.

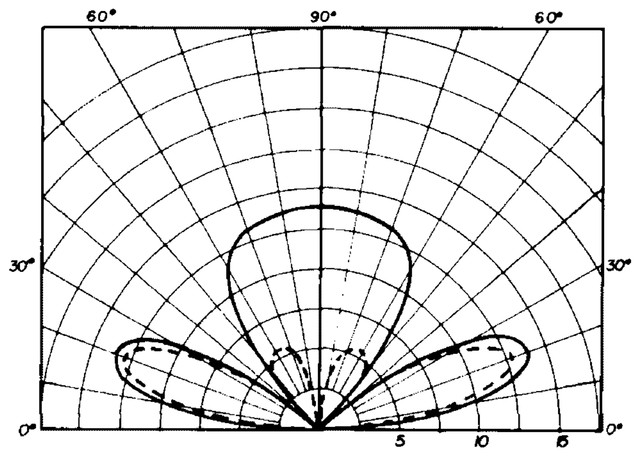

fig. 4. Horizontal loop vs horizontal dipole, both at three-quarter wavelength above ground.

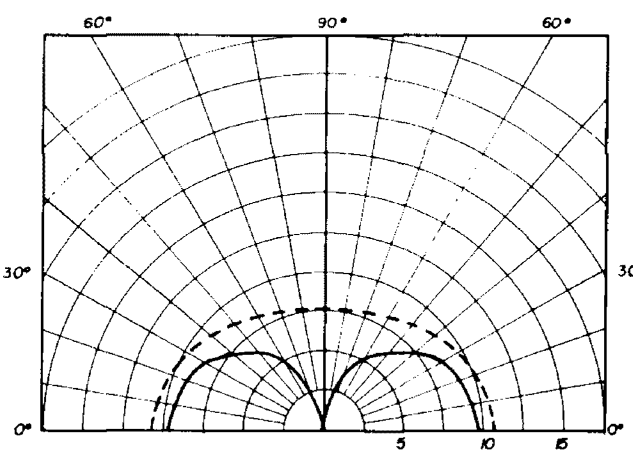

fig. 5. Loop vs linear antenna, both one-eighth wavelength above ground. Both antennas are vertically oriented in this and figures 6, 7, 8 and 9.

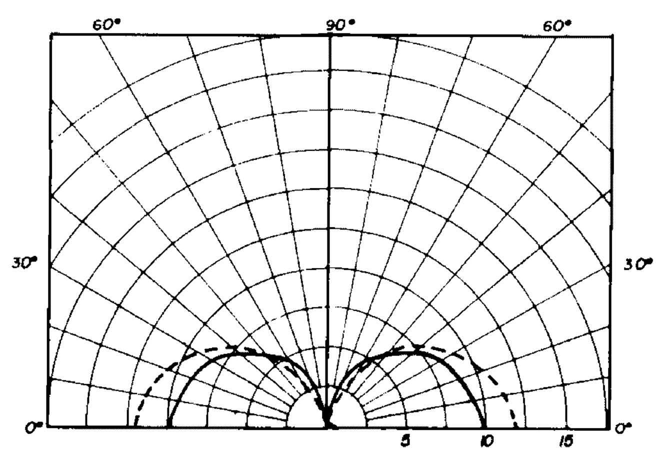

fig. 6. Loop vs linear antenna, both one-quarter wavelength above ground.

shown is that in the plane of the loop. The quantity plotted is the field strength at one meter, and is measured in volts per meter for one watt input to the antenna.* Since actual field-strength units are used, it is possible to compare the performance of the different antennas directly. For example, it has been claimed that the so-called "army loop" is equivalent to a horizontal dipole about one-sixth wavelength high. The reader can pick off a good approximation for each from the charts given here, and can decide for himself how closely this claim approaches reality.

The purpose for which an antenna is intended should be borne in mind when considering alternatives. DX chasing requires low angle radiation, but for a couple-hundred-mile range it is the higher angles that are most useful. Considerable directivity is sometimes an advantage, as when there is heavy interference from a direction other than the desired one. On the other hand, this sometimes requires rotating the antenna to give adequate coverage.

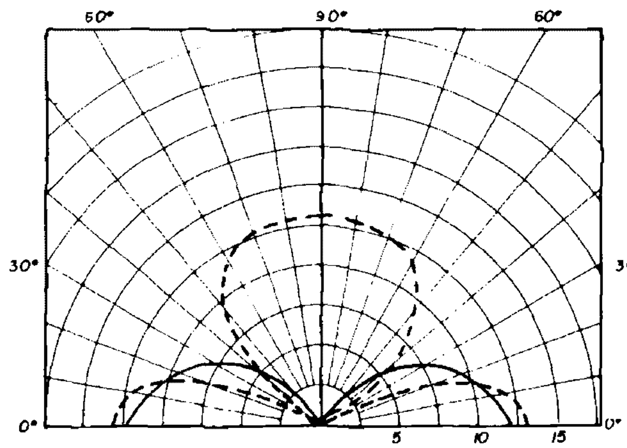

fig. 7. Loop vs linear antenna, both one-half wavelength above ground.

Looking at the charts, it can be seen how the directivity of the antennas plays its role. In the horizontal group, the loop radiates equally in all geographic directions, and as a result its radiation pattern is noticeably smaller than that of the dipole. Even so, for the longer distance contact, the two are very nearly equal. The same cannot be said for short haul work, however. At a height of three-quarter wavelength, for instance, the low angle lobes are almost identical, while at the higher angles the loop responds only feebly compared to the dipole. This might be considered quite a good thing under certain circumstances. If horizontal polarization is desired, the loop is usually easier than a dipole to put up in a densely populated area. And it is precisely in such areas that heavy interference from short range stations is most likely to be encountered.

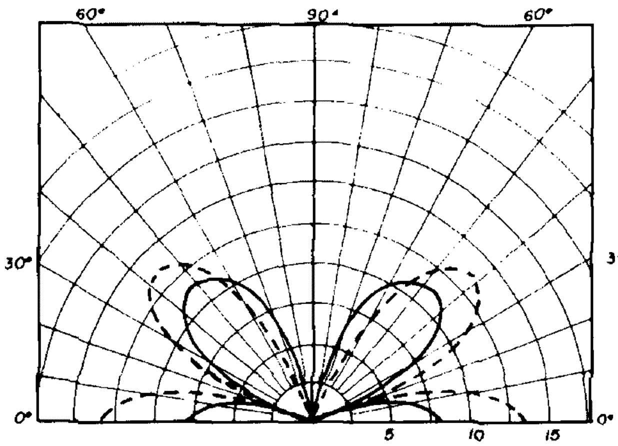

fig. 8. Loop vs linear antenna, both three-quarter wavelength above ground.

In the vertical group, the situation is different. Here it is the loop which is the more directive in the horizontal plane. The result is that it readily outperforms the linear antenna. Of course, this is only in the plane of the loop. Off the side, the signal will be down, just as it is down off the ends of a horizontal dipole. The most marked advantage of the loop for DX work appears when the antenna is between one-quarter and five-eighths wavelength high. See fig. 9. Here, the loop shows a gain of three to four decibels over the linear antenna. This is equivalent to a two-element beam.

*Some texts use as standard the field of one kilowatt at a distance of one mile. The values given in the charts can be converted to this by dividing by 1609 — the distance of one mile in meters — and by multiplying by the square root of the power — 1000 watts. **editor.**

As with all high frequency antennas, radiation at extremely low angles is greatly attenuated by ground absorption. The charts, therefore, are not reliable below nine or ten degrees.

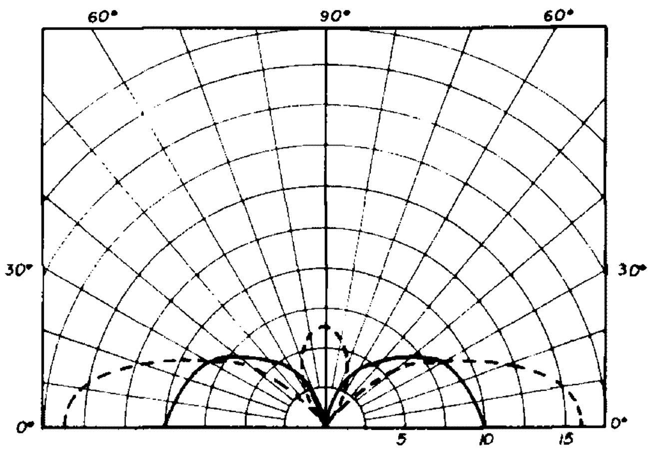

fig. 9. Loop vs linear antenna, both five-sixteenth wavelength above ground. This is the height at which the gain of the loop over the linear antenna is maximum.

losses and efficiency

All the above has been based on the presumption of zero internal losses in the loop — that the loop is composed of a lossless conductor. This may seem to be begging a very important question in view of the admittedly very low radiation resistance of the loop. What interpretation, then, can be placed on such examples? The answer, of course, is that the presence of internal losses reduces the radiated power, and the comparisons to the linear antennas are no longer valid. The directional characteristics, however, should remain much the same. Furthermore, although no antenna can be made truly lossless, this condition can actually be approached in practice.

Let's examine this matter a little further.

First, we need a practical example. Let's compute the radiation resistance of a small loop. The formula for this is given in one form or another in most books on antennas as follows:

$$R_r = 31{,}200\left(\frac{A}{\lambda^2}\right)^2$$

where R_r is the radiation resistance and $\frac{A}{\lambda^2}$ is the area of the loop in square wavelengths. If there are several turns in the loop, the radiation resistance given above is multiplied by the square of the number of turns. However, we are only considering single-loop antennas here.

Let's assume a round loop with a radius of 0.05 wavelength.

$$R_r = 31{,}200(\pi r^2)^2$$

$$R_r = 31{,}200\ (3.14 \times .0025)^2$$

$$R_r = 1.94 \text{ ohms}$$

This corresponds to a forty-meter loop about fourteen feet across.

Suppose we use copper clad down-spouting three inches in diameter for a conductor. To compute the losses in the conductor, we need the resistance — the high-frequency resistance, not the dc resistance. For round, tubular or solid copper conductors, this quantity is given by

$$R_{hf} = \frac{\sqrt{F}}{1000d} \text{ ohms per linear foot}$$

where F is the frequency in megahertz and d is the diameter in inches.[1] If aluminum is used, multiply the result by 1.28, the ratio of the square root of the restivity of aluminum to that of copper.

The high-frequency resistance for our example figures out to be about 0.039 ohm total.

Now, the radiation efficiency of an antenna can be determined by the following formula:

$$\% = 100\ \frac{R_r}{R_r + R_{loss}}$$

Here, this comes out as

$$100\left(\frac{1.94}{1.94 + .039}\right) = 98\%$$

Since it is doubtful that a 2% loss could be measured at any distance, this can be taken as an example of an essentially lossless antenna.

It might be instructive to consider what happens if the conductor is made about fourteen feet across.

Suppose we use copper-clad down-tubing. The high-frequency resistance of this conductor is twelve times that of the three-inch material, or about 0.466 ohms. Yet when this is plugged into the efficiency formula, the result is still 80%. This represents a loss of approximately one decibel.

Now let us assume we wish to use the same size loop and tune it down to the low end of the eighty-meter band. What do we run into in this case? Well, the radiation resistance can be computed by using the formula above. However, we can obtain the same result somewhat more quickly by noting that the radiation resistance is inversely proportional to the fourth power of the wavelength. If the wavelength *is* doubled, the radiation resistance is divided by 2^4, or 16, which gives us a figure of approximately 0.12 ohms for our loop at 3.5 MHz.

Using the quarter inch copper tubing, the efficiency falls to 27%. This sounds rather low, but in point of fact it is roughly thirteen times the efficiency of a bottom loaded eight-foot whip mounted on an automobile, at the same frequency. The three-inch material, however, produces the more desirable efficiency of about 80%.

In actual practice, there are other and less readily-calculated sources of loss in a small loop. One of the worst offenders is the discontinuous joint. Where conductors are merely twisted together, or bolted, the current is not only crowded into a small space, but often must flow through layers of oxide and accumulated dirt. Remember, here we are dealing not in ohms, but in tenths of ohms. When the radiation resistance is very low, as it is in loops, joints – where unavoidable – should be either soldered or welded. Bare conductors should be sprayed with plastic paint to preserve the surface from corrosion. Corners should be rounded where possible. And after attention to all of these things, it is still wise to employ a belt-and-suspenders technique by choosing a somewhat larger conductor size than actually required by the figures.*

ground and radiation resistance

So far, we have only considered the loop's free-space radiation resistance. The *fact that the ground reflection* has such a pronounced effect on the radiation pattern should prepare us for the likelihood

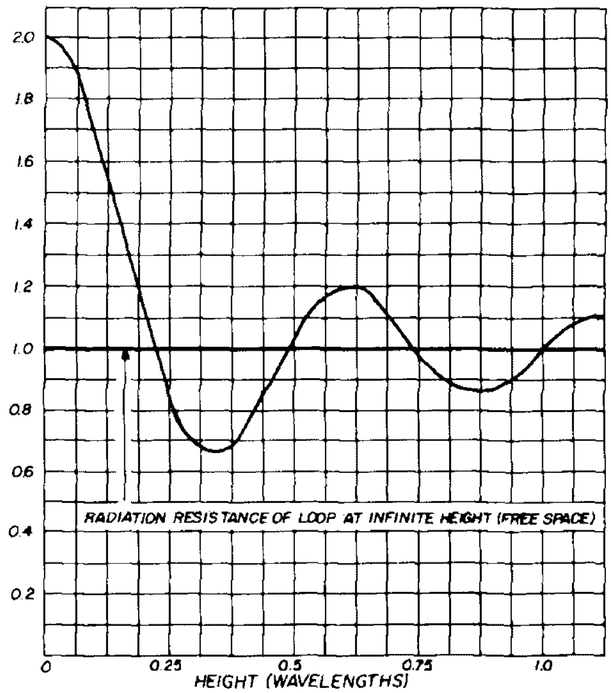

fig. 11. The ratio of the radiation resistance of a small horizontal loop near ground to that of the same loop in free space.

of a similar effect on the radiation resistance. This proves to be the case, and **fig. 10** and **11** tell the story.

What the charts display is not radiation resistance itself, but something more general. It is the *ratio* of the radiation resistance of a loop near ground, to that of the same loop in free space. These charts are applicable to small loops of any size whatever.

It will be observed that a vertical loop has its resistance doubled when very close to ground, while the resistance of the horizontal loop tends to zero at the same

*"Belt and suspenders technique." This term appears to have originated in connection with an eminent civil engineer of some years ago, who was well known for his reliable but extremely conservative structural design. To keep his pants up, this gentleman always wore both belt and suspenders – presumably on the theory that if one should unaccountably fail, *his modesty would still remain intact.* **editor.**

height. From this it is a pretty obvious conclusion that, unless it is to be elevated appreciably, the horizontal loop is pretty well ruled out. It is almost certain to suffer seriously from internal losses — this, even with the largest practicable conductor.

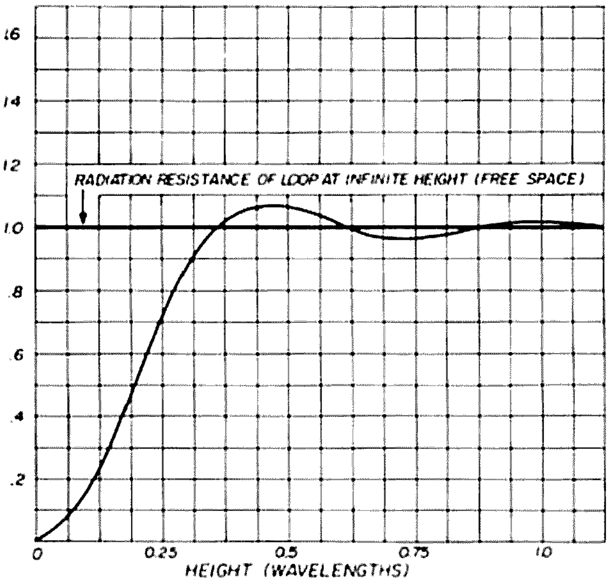

fig. 10. The ratio of the radiation resistance of a small vertical loop near ground to that of the same antenna in free space.

appendix

The directional characteristics of loops near ground are given by the following expressions:

Horizontal loop $E = \sqrt{R_{fs}/R_g}\, 2\left|\cos\alpha \sin(H \sin\alpha)\right|$

Vertical loop $E = \sqrt{R_{fs}/R_g}\, 2\left|\cos(H \sin\alpha)\right|$

where R_{fs} is radiation resistance in free space

R_g is radiation resistance near ground

H is height of antenna above ground in degrees

α is the angle above the horizon.

These are developed from a method called "pattern multiplication," in which the radiation pattern of an array of similar isotropic sources is multiplied by the radiation pattern of the individual real source. Further details may be found in J. Kraus, "Antennas," McGraw-Hill, New York, 1950.

reference

1. F. E. Terman, "Radio Engineer's Handbook," McGraw-Hill, New York, 1943, page 35.

ham radio

an antenna coupler

for the three-band beam

Better harmonic suppression and improved loading of your three-band beam with this coax-to-coax matcher

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There are many types of antenna couplers, and frequently they are designed to serve several purposes by a rearrangement of the components. This article covers a specific design: a coupler for a coax-to-coax match when using a three-band beam. The fact that it works equally well with coax-fed dipoles, and by a change of values it will work equally well on 3.5 and 7 MHz, is beside the point.

You may question the need of such a coupler. Read this article, make a few tests, and I think you will be convinced.

advantages

An antenna coupler is not a magic box. It will not cancel standing waves on the feeder and may do little to prevent radiation from the coax outer shield. It can, however, increase radiated power and attenuate harmonics—both are important items. This coupler, properly used, can do more. It can be used as an output power meter, and it is a first-class tool to use for obtaining an accurate match between antenna and feeder.

The pi-net output tank of a transmitter is intended to match the load resistance of the output tubes to the radiation resistance of the antenna (finally) and at least to provide a match to the coax feeder. In the perfect case, the transmitter will see a perfect load which will draw and dissipate the maximum power available.

The design of a pi-net is not simple. We need to know the exact plate voltage, full-power plate current and the load resistance to be fed before we can calculate the values for the plate inductor and loading capacitor. To obtain maximum power output, therefore, we must operate the tubes at their rated input power and present the correct load resistance.

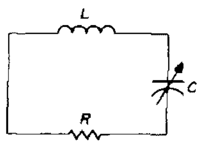
If we have a transmitter designed to operate at X watts into a load of 50 ohms, it will be overloaded if it sees 40 ohms, and will not be loaded to full power with 60 ohms. It must have a load of 50 ohms to give its proper efficiency.

There is not a three-band beam in existence which will provide an exact load of 50 ohms at every frequency on all three bands with its coax attached directly to the pi-net output.

If we use a dummy load of 50 ohms, the transmitter will tune up exactly at any setting and will load to rated input. What we need, primarily, is a device which will at least enable the transmitter to see its optimum load, just as though it were feeding a dummy load. In fact, the antenna should draw exactly the same power as the correct dummy load at any point in any band. Does yours? If it does, you do not need an antenna coupler!

Using 50-ohm coax into a transmitter rated 50 ohms, we do not require any impedance transformation. We do need an unbalanced-to-unbalanced coupler which will cancel any reactance and allow the transmitter to see its proper load. This is not as simple as it sounds, but at least we are able to move in the right direction as a start.

fig. 1. The basic Z-match. At resonance the inductive reactance and capacitive reactance *cancel each* other out.



the z-match

The series-tuned circuit (z-match) is ideal for our purpose. The basic circuit is shown as **fig. 1**. At resonance, the reactances of L and C cancel each other and their net reactance is zero. The internal impedance is therefore equal to the value of the load R.

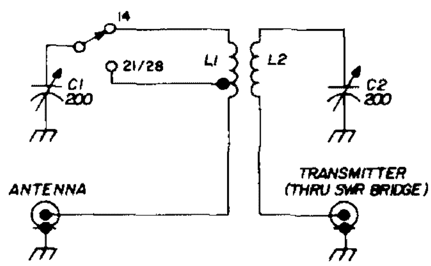
Even if R is reactive (such as a length of coax with a standing wave on it), the reactances form a part of the circuit together with L and C, and can be cancelled by adjusting the values of L and C. By using two such series-tuned circuits, (as in **fig. 2**) we are able to absorb any feeder reactance in circuit L1C1 and present the transmitter with its 50-ohm load.

design

Let us first deal with the design and adjustment of **fig. 2.** L2C2 merely needs to be able to tune to resonance to any frequency between 14 and 29 MHz. This can be done easily with 1-μH inductor and a 200-pF capacitor. There is nothing critical in this part, except that C1 may see rf voltages as high as 300V with a 1-kW input, and L1 may carry as much as 5 amperes. Inductance and capacitor should be capable of handling these values—not a difficult matter.

Circuit L1C1 will need more inductance for 14 MHz. The capacitor may see as high as 1 kV of rf, currents as high as 5 amperes may pass through the inductance, and again, the components must be of appropriate rating. The inductance L1 may need to be around 4 μH total to cover all three bands, and it may need tapping to provide 1 μH for the two higher bands.

The experimental coupler was designed for 400W PEP. The capacitors were both 200-pF maximum, with C2 a good quality receiving-type and C1 of a type used in 50-watt transmitters. The coils were made from a single piece of B&W 1¾-inch diameter, 16 turns-per-inch



L1 7½ turns, no. 18 wire, 1¾-inch diameter, 16 turns per inch (B&W no. 3023), tapped 3½ turns from hot end. Approximately 4 μH

L2 3½ turns, no. 18 wire, 1¾-inch diameter, 16 turns per inch (B&W no. 3023) approximately 1 μH. Note: L1 and L2 are close-coupled on the same piece of coil stock, which altogether measures 11 turns

fig. 2. Schematic of the coax-to-coax antenna matcher and its installation with an swr bridge between the station transmitter and antenna.

stock, 11 turns long. The inductance was merely cut to provide two close-coupled coils of 3½ and 7½ turns. The larger coil coupling make sure that the transmitter is connected to the very *outside* turn of L2 and the antenna is tapped to the end of L1 closest to L2.

You need a dummy load and an swr bridge for initial alignment. On each band, load the transmitter into the dummy load and do not retune the transmitter controls. Insert the tuner between the dummy load and the swr bridge, set switch SW1 to the correct band and adjust C1C2 until the swr bridge (set to read reflected power) shows zero reflected current. Note the settings of C1 and C2 for resetting. As you make these adjustments, you will learn how C1 and C2 function in order to tune the coupler. I had no trouble with the 21- and 28-MHz bands during the basic alignment, but the behavior on 14 MHz should be noted. The amount of L1 in circuit should be varied by tapping the coil, until C1 is about half in. Do not remove the unused turns.

Remove the dummy load and connect the antenna in its place. Begin tuning on 28 MHz. It should be possible to adjust C1C2 so that zero reflected current is obtained on the bridge, although the position of C1 may not be the same as with the dummy load. At this point, mark the 28 MHz settings for C1C2; they will not vary very far over the entire 28-MHz band. Follow the same procedure for 21 MHz, and obtain calibration marks for future routine resetting.

Finally, set up for 14 MHz, and here the tests must be more exhaustive. Try three settings: 14,000, 14,150 and 14,300 kHz. If you obtain zero reflected current at all three, all is well. In the process you may need to vary the amount of L1 in the circuit as this will depend on the amount of reactance presented by the feeder. If the *supposed* resonant frequency of the antenna is known, check at this frequency also.

Experience says that it will be impossible to obtain a perfect match at all points on 14 MHz. However, at the points where matching is nearly correct, adjust the 14-MHz tap on L1 to ensure adequate capacitor movement. Although there may be a point where the bridge will not zero, the swr will not be too great and may be ignored for the moment.

The results of the coupler may now be assessed, both in transmitter loading (freedom from reactances) and behavior on the air. Output power will be higher and harmonic emissions should be attenuated by some 30 dB.

A refinement which you may add is an rf ammeter in series with L2C2. This will read the actual current in the link between the transmitter and the coupler, and output power may be calculated from I^2R. For 50-ohm coax, 50 watts will show 1A; 200 watts, 2A and 1 kW almost 4.5A. The size of the meter will therefore depend on the power to be handled. You will find that, so long as the reflected power is zero and the coax between transmitter and coupler is short (say 6 feet or so), this power calculation will be quite accurate.

matching the antenna

We now come to the reason for a possible inability to obtain a match at some point on the 14 MHz band. The first point we must remember is that the coupler has *not* reduced the standing waves on the feeder, nor done anything to improve the match between antenna and feeder; it has merely cancelled feeder reactances by resonating the feeder.

There is one well-known fact about a resonant feedline (whether it be resonated to a half wave or a multiple of a half wave) and that is that the value of its terminating resistance will be accurately reflected back to the power source. In the case of the three-band beam, with an average radiation resistance of 50 ohms, we see something very near to 50 ohms at L1C1 in most cases. The Z-match will permit a slight impedance transformation by varying C1 against C2, and as a result, we are able to present an accurate 50-ohm match to the transmitter over the whole of the 21- and 28-MHz bands and over quite a large portion of 14 MHz. The impedance transformation of the coupler is, however, limited. You may reach a point where it is no longer possible to obtain sufficient transformation and a true impedance mismatch results.

Let us suppose that the antenna is presenting a radiation resistance of 200 ohms. Even with reactances cancelled, L1C1 sees 200 ohms, but L2C2 requires 50 ohms. The coupler struggles to effect transformation, fails to do it completely, and the transmitter is underloaded.

On the other hand, suppose the radiation resistance falls to 20 ohms. Again, the coupler struggles to transform 20 to 50, but just cannot do the job, and the transmitter is overloaded.

In either of these cases, we have a mismatch which shows up at the swr bridge, and is seen as either an excessive or insufficient current in the rf ammeter. We can, of course, tell whether the antenna is showing a low or a high radiation resistance by the ammeter.

In point of fact, a three-band beam is likely to show a very low radiation resistance at its resonant point. This is usually offset by making the feeders somewhat longer than a half-wave at this frequency in order to obtain some feeder transformation.

Whether it is worth trying to vary the feeder match at the antenna is debatable, because we may upset the match at all other frequencies by doing so. Perhaps we should accept the fact that we have a slight mismatch and let it go at that. However, we should not lose sight of the fact that we have the tool to do the job of matching. Recall that with the coupler, we may terminate the coax in a dummy load instead of the antenna and adjust everything until we have a perfect balance. If we now reconnect the antenna and vary the match, without moving anything else, the swr bridge will tell us when we have made the match.

The important thing here is that the bridge will tell the truth! In the case of a bridge in the main coax line, any slight standing wave on the coax will upset the bridge and any slight change in coax length or in the placement of the bridge along the coax will produce misleading results. We need to isolate the bridge from the main feeder if we are to obtain accuracy—and this is where the coupler comes in.

The routine adjustment of the coupler is simple, provided you made reset marks for C1 and C2. If the capacitors are set to their marks, the transmitter is tuned and loaded for maximum forward current at the bridge (or at the rf ammeter) and C1C2 is then trimmed until there is zero reflected current. As a rule, the bridge should be left to read reverse current, and as the transmitter is moved around in the band, any rise on the meter should be compensated by adjustment of C1.

summary

With the bridge to warn of any rise in swr, you will find that, at any frequency in any band, the transmitter will draw its rated power. Although there will still be some standing waves on the feeders producing feeder losses, these losses are a percentage of the available power and, therefore, any improvement in available power will improve the radiated signal.

Remember, what holds true for transmission also applies to reception. If harmonic radiation is down 30 dB, then the coupler will attenuate incoming signals which are not in the chosen band. When changing bands on the receiver, do not forget to switch the coupler or the band may appear to be dead.

On the subject of harmonics, we should remember that a three-band beam, by its very nature, will accept 28 MHz with the same facility that it accepts 14 MHz; there is no attenuation of the second harmonic of 14 MHz. Moreover, any dipole will accept some power on its harmonics and a three-band affair is likely to radiate the second harmonic of 21 and 28 MHz. With the coupler in use, the antenna is no longer able to load on more than one band at a time because of the filtering action of the two tuned circuits.

acknowledgement

Thanks are due to ZS6YK for his suggestion that *he needed a coupler, for* his suggestion of the rf ammeter and for his painstaking testing of the experimental unit which has now been incorporated into his station.

ham radio

loading the mobile transmitter

Sound information on overcoming the problems frequently faced in loading mobile rigs into the less-than-ideal antenna

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At one time or another, nearly every ham has had trouble loading his mobile rig. The following is a review of some of these problems and what may be done to correct them.

What is proper loading? Proper loading could be defined as the condition under which you are able to obtain satisfactory plate current readings at the settings of the power amplifier tuning and loading controls specified by the manufacturer. For example, the Collins KWM-2 has a section of the final amplifier tuning control marked for the various bands and the load control has a spot marked *50 ohms.* The Heath HW-22 instruction book says loading will be proper when the final amplifier resonates in the middle third of the dial range. If the manufacturer does not specify the approximate locations of these controls, I suggest you connect a 50-ohm dummy load to the transmitter, load the final amplifier to the recommended plate current, and record the dial settings. This should be done on all bands. Whenever the antenna is connected to the rig the final tuning and load controls should be in about the same places as for the dummy antenna. A large deviation means the antenna presents a load too far removed from the optimum non-reactive 50 ohms for proper operation.

impedance

If our mobile antenna does not load like the dummy antenna, or if the transmitter will not tune satisfactorily as recommended by the manufacturer, then the antenna must be adjusted. Further proof of this may be obtained by substituting the 50-ohm dummy load for the antenna at the antenna end of the 50-ohm transmission line. The transmitter should return to normal regardless of the transmission line length. Our job is to

make the antenna termination look like the 50-ohm dummy antenna. This takes real work since the radiation resistance of a base-fed quarter-wave antenna is about 35 ohms when operating over a perfectly conducting ground. Obviously, we rarely have a perfectly conducting ground. In fact, we rarely have the optimum mounting recommended by the antenna manufacturer. Should the optimum ground prevail, there is an interesting chart in an article by WØJF which shows that 50 ohms can be obtained by making the *antenna* slightly longer than for resonance.[1] However, this makes the termination reactive. In other words, we have 50 ohms but it is not like our non-reactive *dummy load*. This will probably show up by our transmitter not tuning properly — particularly the final amplifier will not resonate in the correct spot. To correct this condition, place a small capacitor at the antenna termination to balance out the inductive reactance, and a 50-ohm resistive termination will result. Phil Rand, W1DEM, says he carries along several mica capacitors ranging in value from 50 pF to 0.002 μF for this purpose.[2] He uses the smallest one necessary to provide proper operation. Kayla Bloom Hale, W1EMV, goes Phil one better when she mounts a rotary switch at the base of her mobile antenna and permanently installs capacitors for each band.[3]

swr

I have not mentioned standing waves or reflected power because it can usually be said that if the transmitter loads properly then the swr is within tolerance. But what is tolerance? Many of our transmitters — particularly high-power linear amplifiers — are designed for no more than a 2:1 swr. This, then, determines the tolerance. W1DBM says a 5:1 swr is hardly worth worrying about on 4 MHz as far as loss in concerned, but many transmitters are not designed to compensate for this condition, therefore it is well to use a meter to get the swr down as low as possible by using the capacitors and adjusting the antenna. He also points out that with a higher swr, transmission line lengths become important. The reason for this is that the impedance varies with the line length. This impedance is determined by the voltage and current at each particular spot on the line. For example, in a properly terminated 50-ohm line, we could have 100 volts and about 2 amperes of rf at the transmitter running 200 watts output. At the antenna end we would have the same thing, assuming no loss in the line. However, if the antenna were so adjusted that a 12.5-ohm impedance was presented to the line, we would have a 4 to 1 mismatch, and we would also have a 4:1 swr. Theoretically, depending on the line length, this could result in an impedance between 12:5 and 200 ohms, resistive, capacitive or inductive, at the transmitter end. It might be possible to come up with a length of line that would be close enough to 50 ohms, at some specific frequency, for the transmitter to load satisfactorily. However, more than likely, it would fall somewhere between the maximum and minimum values at a point where the transmitter would not load. If we happened to have a half wavelength of line (electrically), then the impedance presented to the transmitter would be the same as at the antenna. Such a length is not normally considered desirable.

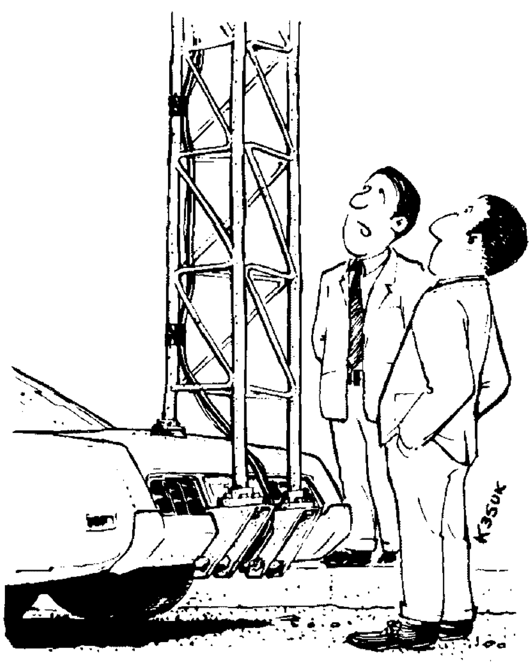
reflected power

Normally, reflected power is not lost. If the transmitter is capable of loading properly into a *line of high* swr, it will be found that nearly the same power is delivered to the mismatched load as to the matched load. Carl C. Drumeller, W5JJ, demonstrated this when he substituted a load of one quarter the terminating resistance, resulting in a 4 to 1 mismatch.[4] When the transmitter was loaded to the same power input for both terminations, the total wattage in the load was very nearly the same. The current in the lower resistance had nearly doubled which, when applied to the current squared times resistance formula, came up with the same power. This again bears out the fact that if the transmitter

has the capability of loading into wide variations of load impedances, then the swr is not so important. What does happen with high swr is a tendency for loss in the transmission line caused by heat or dielectric loss at the points where high currents or voltages occur. Many hams use antenna tuners to tune a line so it will present 50 ohms to the transmitter. This is an excellent device. However, only that part of the line between the tuning device and the transmitter is flat. This system is used frequently with a balanced line to provide all-band operation from one antenna. Depending on the length of the antenna and the frequency, the swr on the line between the antenna and the tuning unit can be extremely high. The fact that these antennas make excellent radiators will again show that, even under these conditions, the indicated reflected power in that portion of the line between the antenna and the tuner is not lost. W5JJ says this about reflected power in his experiments: "In each instance, this power is fictitious, properly measurable in terms of 'volt-amperes reactive' instead of true 'or work-producing' watts."

fixed operation

Most mobile antennas, 10 meters and above, approximate an electrical quarter wave. You can generally say that an antenna less than a quarter wave physically is less efficient than a full quarter wave. This can be dramatically illustrated when going from a 75-meter loaded whip to a full quarter-wave wire of about 60 feet. Even a low 60-foot wire will be found to be far superior. This is no reflection on the mobile antenna design; an eight foot antenna just cannot be expected to compete with one 60 feet long on 75 meters. Also, the maximum radiating portion of an antenna is considered to be that part which has the highest current. With a center-loaded whip, this occurs between the base and the loading coil. It is unfortunate that this part of the antenna is frequently shielded by the car body. When the car is parked or when operating from a trailer, for example, it is possible to connect a 60 foot wire (length adjusted for best swr or optimum loading) at the base of a mobile whip for 75 meters and place a resonator for another band on the whip. This gives you a two band system. Other combinations can be worked out such as 75- and 40-meter quarter wave wires and a 20-meter resonator for three bands. However, contrary to the article by W3HTF, these wires must be connected at the bottom of the whip.[5]



"I'll betcha a steak dinner that he's not married."

summary

In conclusion we might summarize as follows:

1. Know where the final amplifier tuning and loading control should be for the proper plate current recommended by the manufacturer when the transmitter is fully loaded to a 50-ohm resistive load.

2. If possible use a length of transmission line that is not a multiple of an electrical half wave at the frequencies most used.

3. Adjust the antenna so that the transmitter will load according to manufacturer's specifications. Use an swr meter to assist you, as lowest

reflected power and normal transmitter tuning and loading will usually occur at the same place. If satisfactory results can not be obtained add capacitors at the antenna base and retune the resonator or loading coil.

4. Remember that proper transmitter operation is what we are after. It is conceivable that if the transmitter loads and tunes from a widely divergent spot from optimum, some of the power being radiated could be harmonics. This would make the swr meter read incorrectly since it is *not* a frequency-sensitive device.

5. Once you have your transmitter operating properly, do not be influenced to make a change because someone else uses a different length of transmission line! Remember, he has tuned his antenna to match his particular line length and antenna installation. Merely duplicating his line length will probably not help you unless you go through the complete retuning of your antenna system.

6. Mobile quarter-wave antennas are designed to be operated with the base on the car body or close to it. Most verticals of all types — with and without traps — are made to operate with radials or on the ground. Placing the mobile antenna very far above the vehicle on a pole will make it very difficult to tune unless you want to attach radials at the base of the actual antenna.

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ham radio

measuring coaxial-line loss with a reflectometer

Here is a simple technique using an ordinary reflectometer and the station transmitter to determine losses of new and used coaxial lines

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Coaxial lines deteriorate over a period of time, especially when they are exposed to the weather. Periodic checks should be made to determine whether or not the line attenuation has increased to the point where the coax line needs replacement. This is especially true of vhf where line losses normally run higher and where line losses drastically affect both transmitting and receiving. An increase in coaxial-line loss results in a lower vswr at the transmitter when the reflected voltage is reduced by the line attenuation, so the amateur is often unaware of increasing line losses until he starts wondering why he hasn't been hearing and working those DX stations.

The coax feedline must be disconnected from the antenna to make this attenuation measurement, and some provision must be made for shorting the inner conductor to the shield at the end of the line which previously connected to the antenna. A coax connector at the end of the line and a coaxial short are desirable, especially at vhf, but careful soldering of the center conductor to the braid with as short a lead as possible will suffice.

Make several measurements with an

ohmmeter before making the reflectometer measurement. This assures that the coax is not shorted or has an open shield. A short or open circuit condition indicates a bad coaxial line which would not be detected by the reflectometer test. The coax should be immediately discarded if it fails either of these tests. To check for shorts, leave the far end of the coaxial line (antenna end) open and connect an ohmmeter, set to the highest resistance scale, between the center conductor and the shield of the coax.

An open circuit (infinite resistance) should be indicated on the ohmmeter. If the coax passes this test a coaxial short should be placed at the end of the line, and the resistance of the shield and center conductor measured with the ohmmeter set to the lowest resistance scale. This resistance will vary with the type of coax used and the length of the line, but for practical purposes the resistance will be less than one ohm for lengths under 100 feet. An open-circuit or high-resistance indicates that the shield has deteriorated or that the center conductor is open, making the coax useless.

If the coax passes the short and continuity tests, proceed to check the actual loss in the coax at radio frequencies using a reflectometer in conjunction with a low-power rf source connected as in the block diagram, fig. 1. Since losses increase with frequency, the coax should be checked at the highest frequency at which the coax is to be used. The low-power rf source may be a low-power transmitter (5 to 25 watts) loosely coupled to the reflectometer, or the driver stage of a higher powered transmitter link coupled into the reflectometer with a small coupling loop at the end of a short length of coax. Link coupling into the driver stage of a 100-watt two-meter transmitter worked very well for measurements at my station.

Equations:

1. p = Vr/Vf
2. dB = 20 log Vf/Vr
3. Vf/Vr = 1/p
4. dB = 20 log 1/p
5. p = (1 - vswr)/(1 + vswr)
6. vswr = (1 + P)/(1 - p)

In the forward position the reflectometer will read the forward or input voltage to the coax (Vf). This should be adjusted for a full-scale reading, making sure that the far end of the coaxial line is shorted. In the reverse position, the reflectometer will read the reflected voltage (Vr). This reflected voltage is the input voltage attenuated by the coaxial loss, reflected at the shorted end of the coax, and attenuated again by the coaxial loss between the short and the reflectometer.

The ratio of the reflected voltage to the forward voltage is the reflection coefficient (p) as noted in the following equation:

$$p = \frac{Vr}{Vf} \qquad (1)$$

The ratio of the forward voltage to the reflected voltage is used in **equation 2** to determine the attenuation in decibels. **Equation 1** may be rewritten to express the ratio of the forward voltage to the reflected voltage as the reciprocal of the reflection coefficient as noted in **equation 3**. Substituting 1/p for Vf/Vr in **equation 2**, we now have the attenuation of the coaxial cable in terms of the reflection as indicated in **equation 4**.

Thus, the loss in dB can readily be determined by measuring the reflection coefficient of the shorted coaxial line. Remember that this represents the loss going down the line as well as coming

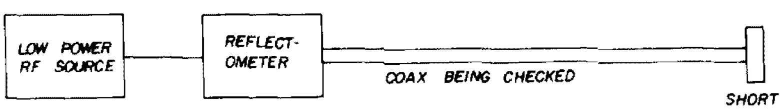


fig. 1. Block diagram of the test setup for determining coaxial line losses with a reflectometer.

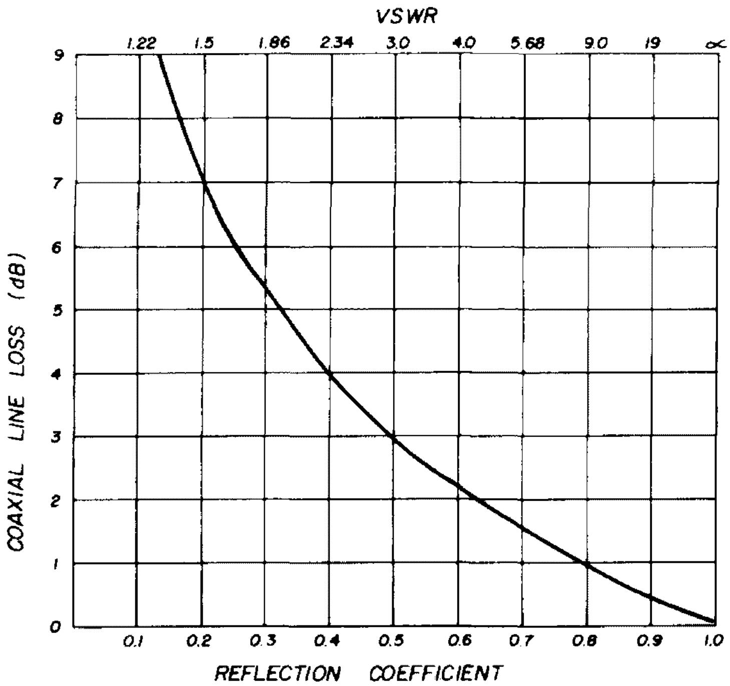


fig. 2. Coaxial line loss is read in dB at left at the point of intersection of the curve and the reflection coefficient.

back, thus the attenuation figure must be divided by 2 to find the actual coax loss. If the reflectometer meter scale reads vswr directly, use **equation 5** to determine the reflection coefficient and proceed as before.

example

As an example, consider a coax line which presents a reflection coefficient of 0.5 when shorted on the end of the line. Using **equation 6**, this represents a standing-wave ratio of 3:1. Using **equation 3**, the ratio Vf/Vr is equal to 2. The two-way coax attenuation is next determined using **equation 2**.

$$dB = 20 \log 2$$

Since the log of 2 = 0.3, the attenuation is 6 dB, and the actual line attenuation is 1/2 of this value, or 3 dB. This may also be determined from the graph by drawing a vertical line at p = 0.5 on the lower horizontal scale denoting reflection coefficient. Note that this line also crosses the upper vswr scale at 3.0:1. Draw a horizontal line where this vertical line crosses the curve and the point where it intersects the left vertical scale indicates the actual coax line attenuation, 3 dB.

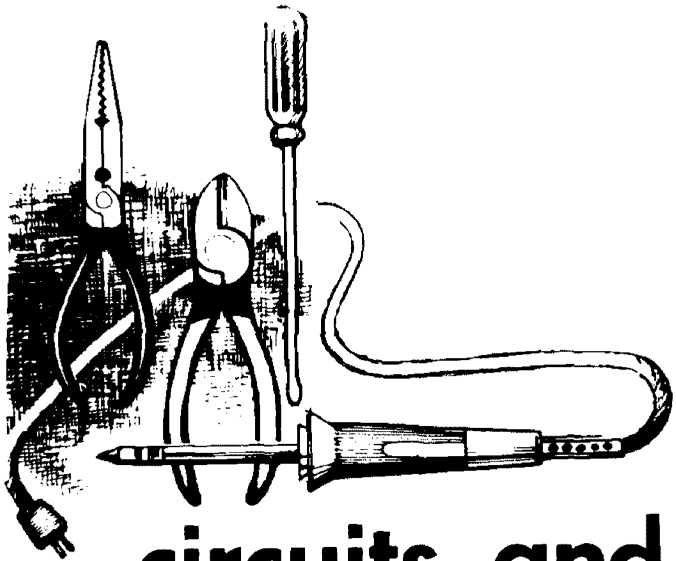
After the coax loss has been measured, the normal loss should be determined to compare with the actual loss measured. **Table 1** lists the attenuation loss in dB for several types of coax at several frequencies. Remember that the attenuation is proportional to the length of coaxial line used and increases with frequency. The attenuation of a 50-foot length of coax will be one half the attenuation of a 100-foot length.

table 1. Coax loss per 100-foot length (dB).

	28 MHz	50 MHz	144 MHz	220 MHz
RG-8/U	1.1	1.33	2.8	3.5
RG-11/U	1.3	1.59	2.85	3.25
RG-58/U	2.5	3.13	5.8	6.9
RG-59/U	2.0	2.4	4.25	4.85

Measurements should be made on several new lengths of coaxial line to properly check out the test circuit and technique. If the results are in question, the coaxial short may be placed at the reflectometer output, where the reflection should be 100%, giving a reflection coefficient of 1. A reading less than this indicates a need for reflectometer adjustment. With reasonable care, the attenuation of coaxial lines of 50 feet or more may be determined with fairly good accuracy.

ham radio



circuits and techniques

ed noll, W3FQJ

antenna potpourri

Here is still another configuration for those of you who have become enthusiastic triangle experimenters. This triangle is open at the top (apex) to attain a low-impedance feed point at the center of the base, **fig. 1**. By so doing, the antenna is divided into two three-quarter wavelength segments and a resultant low-impedance feed. In the broadside direction there is a gain improvement over the full-wave closed triangle. The antenna impedance is about 200 ohms at resonance, and a broadband match can be attained with a 4-to-1 balun. A tee-network tuner also functions well. In the latter case a 40-meter version of the three-half wavelength triangle also performs well on 15 meters where it can be matched to operate as a nine-half wavelength open triangle. I recently erected a 40-meter version. (**fig. 2**) of an antenna I had previously checked successfully on 10, 15 and 20 meters. The basic three-half wavelength equation would be:

$$\text{length in feet} = \frac{1476}{f(\text{MHz})} = 206'$$

Therefore, approximate length for resonance near the center of 40 meters would be 206 feet (two 102-foot segments). In practice the band-center resonance was obtained by clipping back about 5 feet on each side. This corresponds to a shortening of about 5%. The amount of shortening will vary with the height of the triangle's base above ground and the apex angle of the triangle. First cut for formula length and then shorten. The swr figures, when using a 4-to-1 W2AU balun, are also given in fig. 2.

triangle advantages

The advantages of triangle antennas are good performance at low cost, single mast erection, transmission line need only be run to the base of the triangle and not up the mast, feed ends of the line are readily available for trimming and matching, directional pattern can be shifted from ground level simply by swinging the triangle corners around to other pairs of metal fence posts that act as supports for the base of the triangle and the vertical angle can be changed easily.

tilt angle experiments

The tilt angle of the triangle plane can be changed conveniently by moving the metal fence posts which support the base corners of the triangle. The vertical angle is lowered by moving the base of the triangle toward the preferred direction. In tests, the most favorable tilt for late-night reception from Europe was about 15°.

Test procedure was unique and

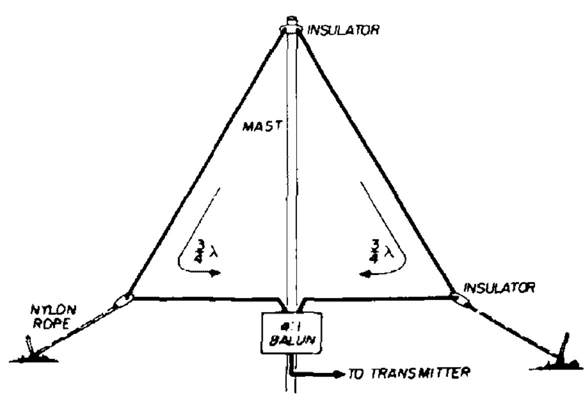


fig. 1. One-and-a-half-wavelength triangle antenna.

economical—I used those pesky foreign broadcast stations on 40 meters. A comparison antenna was a full-wave 40-meter wire, sloped toward Europe. These tests made one significant impression: the vertical angle of the receiving antenna has a definite influence on the signal delivered to the receiver input. Furthermore, that preferred angle is not a constant. In one test session with the triangle in a vertical position, the early evening reception on the triangle was better than the sloping wire. Much later the same night, stronger signals were being delivered by the sloping wire.

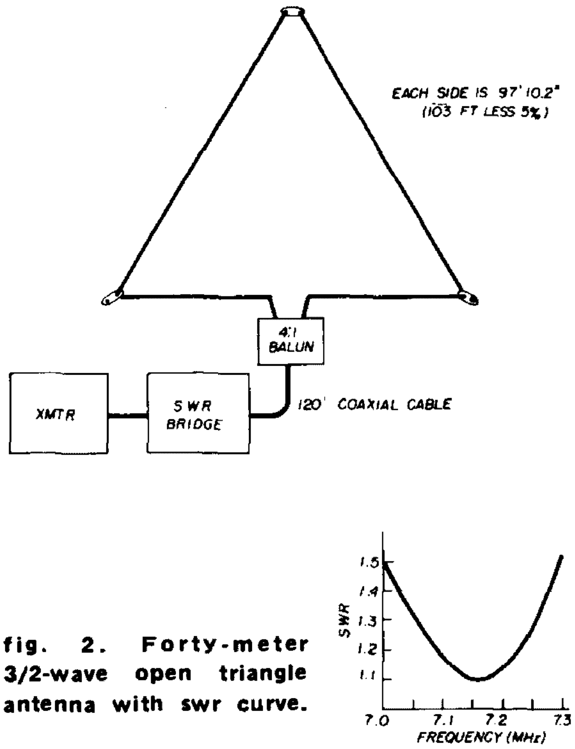


fig. 2. Forty-meter 3/2-wave open triangle antenna with swr curve.

All of this might indicate that even in the case of the popular sloping dipole used widely for low-band DXing, a remote means of changing the height of the low side of the dipole may be of definite help in establishing a favorable angle for the propagation conditions of the moment. The influence on 80 and 160 meter low-angle performance could be important. You can gain experience on 40 meters by using the foreign broadcast stations as test signals.

Tilting a triangle in the preferred direction lowers the vertical angle. At the same time the vertical angle is raised to the rear.

more low angle

The subject of low-angle radiation was treated at length last year by Pat Hawker, G3VA.[1] He mentioned that during poor conditions on the North Atlantic path the vertical angle of radiation should preferably be below 2.5°. Certainly these conditions will prevail during minimum sunspot and it may be that the very popular beam antenna of today will not be the one that does the best job during minimum sunspot years. Ground interference can have an adverse affect on the vertical patterns of such antennas. Also important is the matter of obtaining effective height; that is, at least one wavelength above ground.

Pat also mentioned the attractive characteristics of sloping antennas and the use of verticals to obtain low-angle radiation. Dipoles, vees and rhombics can be sloped to obtain low-angle radiation at moderate antenna height.

The article states that the sloping-V antenna, fig. 3, has been neglected. Only a single high mast is needed with the antenna ends dropping to near ground level. A good unidirectional pattern and broadband performance is obtained using 600-ohm terminating resistors. Transmission line can be tapered from a high of about 800 ohms to the several hundred ohms of one of the common types of

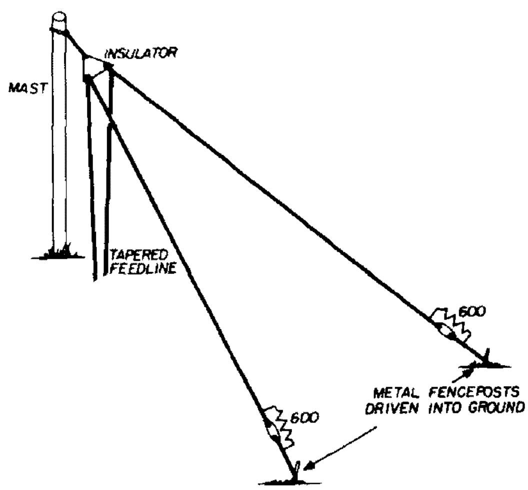


fig. 3. Sloping V-beam antenna.

open-wire transmission line. An alternative is to taper the line down to about 200 ohms and use a 4-to-1 balun to match and transform the antenna system to a coaxial transmission line. One of the favorable characteristics of sloping rhombics and sloping vees often overlooked is their broadbandedness as compared to strictly horizontal versions. This is attractive in terms of multiband amateur operations.

umbrella antenna

An interesting article by James K. Palmer concerned the umbrella antenna, an elaboration of the inverted V.[2] Inverted V segments cut to various frequencies are spaced equidistantly about the single mast of the antenna system. A typical model operating between 4 and 30 megahertz with a maximum swr of only 1.5 to 1 consisted of 10 dipole antennas (20 elements) cut for frequencies of 4, 7, 10, 15, 18, 21, 23, 27, 29 and 30 megahertz.

I have used a similar maypole construction successfully on the amateur bands, although not as many elements were used.[3] One was a 15-, 40-, 80-meter dipole combination for Novices and the other was a 20-40-80 combination for

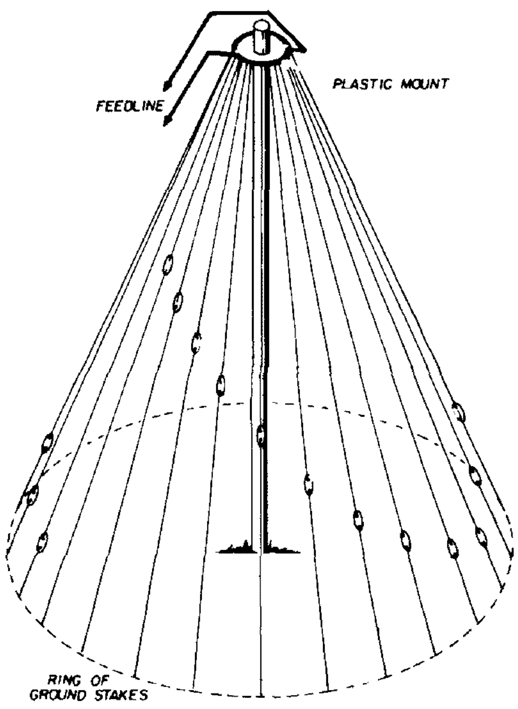


fig. 4. Umbrella antenna.

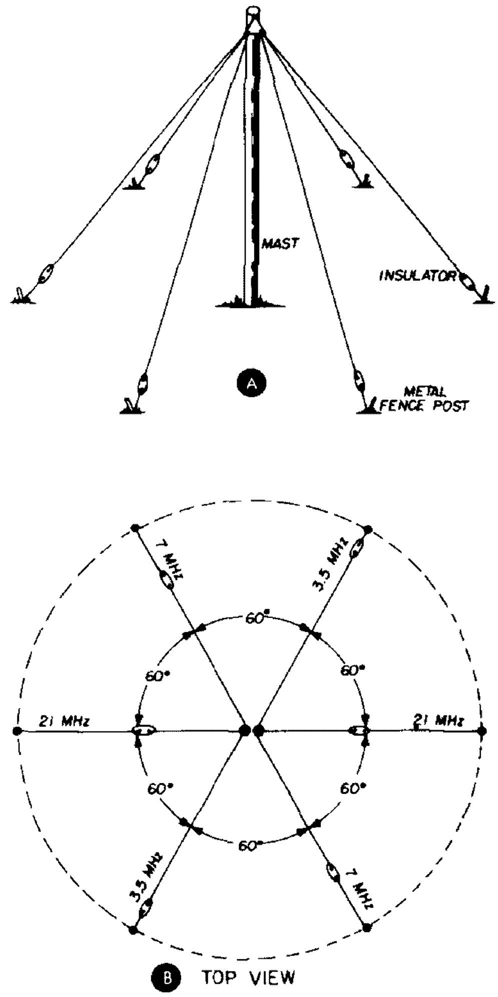


fig. 5. Construction of the Maypole antenna.

side-band operation, fig. 5. However, there is no reason that the idea cannot be expanded into a good broadband, all-band antenna with low standing wave ratio.

yagi arraying

One of the most exciting possibilities for amateur operations can be anticipated from the very fine and detailed Yagi article by James Emerson.[4] One of the most trying problems in providing a high-quality community antenna television system (CATV) is co-channel and adjacent channel interference. These are also some of the biggest problems in amateur radio reception.

Most CATV receiving stations use Yagi or log-periodic antennas. They often use them in pairs or bays to increase receive gain. Sometimes the pairs or bays are appropriately phased to reduce co-channel or *adjacent channel* interference.

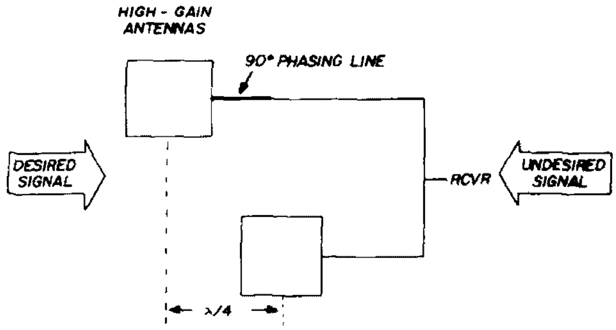


fig. 6. Reducing undesired signal pickup by properly phasing high-gain antennas.

One of the most obvious results of the technique would be the reduction of the signal level of a more distant television station on the same frequency that enters the back of the antenna. With the use of two phased antennas there results a considerable improvement in the front-to-back ratio above the basic front-to-back of the basic antenna configuration. This possibility is shown in fig. 6. Note that one antenna (a Yagi or other type) is positioned 90° ahead of the other. Furthermore, there is a 90° phasing line attached to this antenna before it reaches the combining receptacle. The net result is that a signal arriving from the back introduces a signal into the two Yagi's which cancel out at the combining point. The use of two Yagis, so phased, can do a fine job in cutting down the back pickup (pesky short-skip interference).

It is indeed fascinating that a proper two-Yagi combination can even cut back on interference arriving from the front provided that its angle of arrival is more than 5° away from the angle of the desired signal, fig. 7. Of course, a

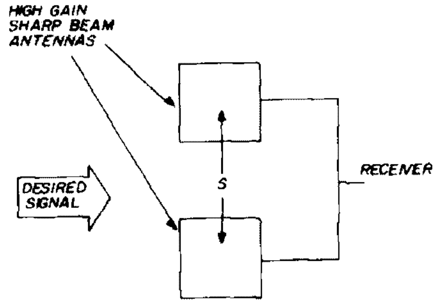


fig. 7. Reducing signal pickup from the front and sides of the antenna. Distance S determines the angle of the nulls.

basically high-beam and sharp-beam Yagi must be employed. Two such Yagis are mounted side by side (broadside). Proper adjustment of the spacing between the two Yagis can be used to locate the pattern null in the direction of the undesired-signal angle. For example, two such sharp beam Yagis separated by one wavelength will produce sharp nulls at 30 and 330 degrees.

Calling all antenna experimenters! What about the possibility of two separately rotatable Yagis (or perhaps three mounted in a triangle) as an effective anti-interference arrangement? In the simplest arrangement perhaps only one

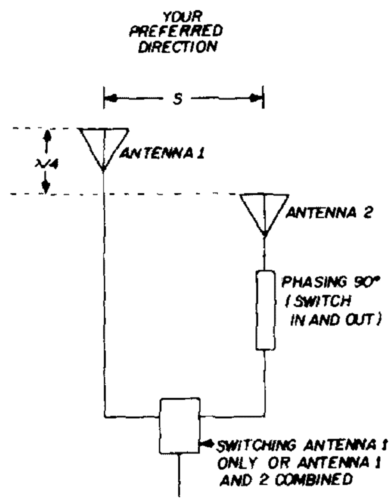


fig. 8. An amateur-band ultimate. Choose distance S wisely for your area.

need be used as a transmit antenna. The second one could be switched in, fig. 8, on receive only. It would be separately rotatable and would be rotated to obtain a null on an interfering station.

references

1. Pat Hawker, G3VA, "Technical Topics," *Radio Communications*, April, 1971, pages 263-265.
2. James K. Palmer, "The Umbrella Antenna: An Inexpensive Broadband Omnidirectional Antenna," *Communication News*, June, 1971, page 52.
3. Edward M. Noll, W3FQJ, "73 Dipole and Long-Wire Antennas," Howard Sams, Indianapolis, Indiana, 1971, pages 23-27.
4. James Emerson, "Arraying Yagi Antennas For Positive Results," *Broadcast Engineering*, May, 1971, page 32.

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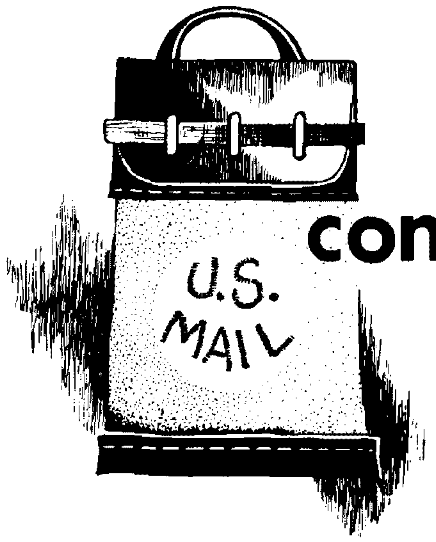


comments

triangle antennas

Dear HR:

I have constructed two of these antennas, one for 20 meters and one for 40 meters. The results on 40 meters have been outstanding, both on receive (low noise) and transmitting. On-the-air tests compared to an inverted vee at the same height ran from 1½-S units to *quote,* "Just like turning on a linear."

I wanted to try this antenna on 75-meter ssb, but since I didn't have a pole high enough or a city lot wide enough, I tried adding to the 40-meter triangle with very good results. See **fig. 1**.

By clipping on two wires approximately 40-feet long at the corners of the triangle and adjusting the length, it tuned up on 3925 kHz. To return to 40 meters, I just unclipped the wires. These 40-foot wires could not be placed in the same plane as the antenna (due to space limitations), but were connected at the corners with an angle of slightly more than 90°.

The wires were connected at a height of ten feet at the triangle corners and six feet at the far ends. For a short test the wires were stretched in the same plane as the triangle and no difference was detected. The swr on both 40 meters and 80 meters was the same, 1.05:1; the antenna was broadly resonant between 3800 and 4000 kHz. Relays could be used or maybe traps *(horrors)*, to make an excellent two-band antenna.

I used the antenna formula $(984/f_o) = \lambda$ and ended up too high in frequency and outside the band (using number-14 enameled wire). I found that $(995/f_o = \lambda$ came out better in my case. Why? Could it be the plastic covered wire in your case and the enameled wire in my case?

L. Showalter, W6KIW
Petaluma, California

I've always been impressed with the results of the triangle. Your manner of obtaining 80-meter resonance is quite novel, and I hope to give it a try. I want to add that the triangle need not be equilateral (three equal-length sides), and you probably could erect an 80-meter version with your present pole.

Three factors affect resonance and impedance. These are apex angle, height above ground in wavelength and any beneath-ground radial system you may use with it. Like an inverted-vee, more length is needed to resonate on a given frequency as the apex angle is decreased. It has been my experience with an inverted-vee that the best low-angle DX results are obtained off the wire ends especially when legs were greater than λ/4

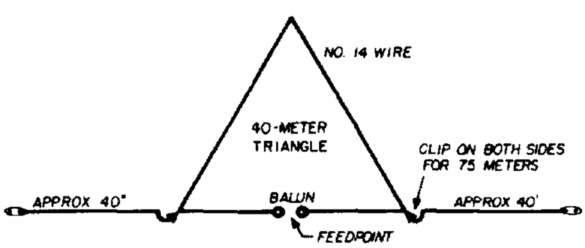


fig. 1. W6KIW's 75- and 40-meter triangle.

long. However, with triangles, the best results are obtained broadside for short and long skip.

It has been a fascinating configuration for me and there is much to be learned. I hope you will continue your experiments and keep us informed.

Ed Noll, W3FQJ

laser communication

Dear HR:

It is encouraging that some readers have expressed interest in laser communication. I have taken out some time for optical experiments that have resulted in a first *two-way laser QSO*! It took place on the evening of 12 November 1971 over a path of 3.8-miles—line of sight—using both a 0.5 mW helium-neon and two 7W gallium arsenide heterostructure laser diodes. Voice communication was attempted. However, we had to resort to keying the gallium arsenide laser units on-and-off because the pulse-fm detector circuits did not detect the narrowband fm satisfactorily.

Ralph Campbell, W4KAE

safer grounding

Dear HR:

I was surprised to read the letter by Keith Olson, W7FS, which appeared in the May, 1971 issue. The driving of ground rods and finding a location with suitable ground conductivity has always been a problem, but why risk your own or your neighbors' health in the process? Mr. Olson recommends filling the pipe with copper sulfate ($CuSO_4$) once it has been driven down to reach the ground-water level. Anyone who has never worked with copper sulfate might think this to be a good practice, since it has copper in its name, and copper is a good conductor. Copper sulfate, however, is very poisonous and caution should be observed whenever it is used.

If you have ever read the ingredients listed on the sides of insect spray powders, you have found that many of them contain a compound known as Blue Viriol or copper sulfate, for they are one in the same. One of the most common sprays is the agricultural one known as *Bordeaux*. If this substance is allowed to enter the ground water supply as Mr. Olson suggests, the results could be fatal. In some areas where city water is not available and your source of water is the well in the back yard or under the house, by using copper sulfate you run a very high risk of contaminating your own or your neighbors' water supply.

As a replacement for the copper sulfate, I recommend good old table salt, also known as sodium chloride. In solution, it is a better conductor and will thus make a better ground. It is also cheaper, easier to find and not nearly so dangerous.

Dennis Recla, WA5KTC
Garland, Texas

threshoid-gate/limiter addition

Dear HR:

A recent article by WB2VXR* points out that the signal distortion introduced by threshold gating may be reduced by filtering the gated signal. Applying a simple W4NVK filter† to the threshold-gate/limiter circuit presented in the January, 1972 issue of *ham radio* (page 46) was found to give a smoother tone and to place less reliance on pre-gating Q to get an acceptable note. The addition of an 88-mH toroid and a suitable capacitor (0.25 μF in our case) in series with the 8-ohm headset proved to be a worthwhile addition. The Q of this circuit is high enough to reduce distortion, but low enough to avoid ringing problems.

John J. Duda, W2ELV
Geneseo, New York

*C. Andes, WB2VXR, Threshold Detectors in a CW Audio Filter, *QST*, December, 1971, page 20.

†E. Dusina, W4NVK, The Simplest Audio Filter, *ham radio*, October, 1970, page 44.

rome tourist guide

Dear HR:

Since I work in Rome as a tourist guide for American tourists, if American hams come to Rome I can show them around in my car at favorable rates. However, they must notify me well in advance, particularly if they come to Rome during the high season from May through October.

Di Pietro Corradino, I1DP
Via Pandosia 43
00183 Roma, Italy
Tel: **756-7918**

phase-locked loop RTTY terminal unit

Dear HR:

In the phase-locked loop RTTY terminal unit described *in the January*, 1972 issue of *ham radio*, a simple R-C charging network is used for autostart detection. The positive-going pulses that occur on space are sampled from the output of the detector card. When enough pulses have arrived to develop 1.6 volts across C204, the Schmitt trigger fires allowing current to flow to the Magtrac.* The Magtrac triggers the SCRs on, turning on the machine motor.

This system works extremely well for on-the-air QSOs and attended operation on some of the MARS RTTY nets. However, it has one drawback: If a CW signal appears on frequency anywhere from the normal VCO resting frequency to the upper capture range limit, the phase-lock loop detector decodes it as a shifted signal. This was quite apparent when I tried monitoring the 14.075, 7117.5 and 3637.5 MHz narrow-shift autostart nets; I would get six inches of garble on the paper from false autostarts caused by a CW signal in the upper half of the phase-lock loop capture range.

The "blank" key provides the greatest space duration of any 60-wpm, five-level RTTY character; the *start* bit plus the five code bits have a total space duration of 132 ms. Thus, any signal with a space condition greater than 132 milliseconds is not RTTY. Since a normal CW dash is longer than this, by adding a pulse width discriminator to the unit, lockout can be set up on all space signals longer than 132 ms.

In addition, a memory or holding function is necessary to keep the terminal unit output in mark-hold between dashes. This is done with an R-C circuit that has a fast attack and slow release. Another function inhibits any further information from reaching the autostart; if the machine motor is on, it drops out during the normal motor delay time; if the motor is off, it is prevented from turning on. With these additions I had an autostart circuit that was fairly foolproof.

Every time a CW signal is detected, the mark-hold circuit activates for 3.5 seconds. After the last CW character having a dash is sent, the terminal unit returns to an "active" or ready-to-print state after the final delay of 3.5 seconds. This anti-CW function is an absolute must for *unattended* autostart operation. In addition, a SELCAL I output is provided for those operators who use the original SELCAL I electronic calling device. In addition to the normal *force-on* provided at the motor control section, a *force-off* has also been provided to allow the SELCAL I to have absolute control when the teleprinter is in the receive mode.

For complete information on this new, improved phase-locked loop terminal unit, please write to me, Box W, Ham Radio Magazine, Greenville, N.H. 03048.

Ed Webb, W4FQM

eliminating the matrix in the automatic fist follower

Dear HR:

The article on the Automatic Fist Follower in the November 1971 issue becomes vague when it undertakes to explain, with the aid of **fig. 13**, the operation of the translator. Even if the

additional 47 circuits in the diode matrix, indicated by **fig.** 13, contain only seven diodes each, the total would be more than 340. They are probably what caused the authors to say, in commenting on the diode matrix board, ". . . it's awkward, takes a lot of room, and we wish it wasn't there."

The matrix and its associated ASCII generator are not easy to understand, especially in view of the statement that the ASCII generator is nothing more than six flip flops connected to count from zero to 63 and repeat. For instance, only five lines are shown coming down to the diode matrix with the letter X. How would the translator tell the difference between a dot and a dash on only five lines? And, five lines can only count to 31 when, obviously, 48 counts are needed. On the other hand if, for instance, ten lines are used instead of five—that is, five from the dot register and five from the dash register—and the ASCII flip flops were to count through 63 for each set of lines for each Morse code character, then, although adding diodes, the logic would be easier to understand. *With only the five lines* there are obviously processes in operation not readily apparent from the figure.

The really hard work in designing the machine, for which the designers should be given no end of credit, was done when the dots were separated from the dashes so they could be placed in separate registers. After that it was only necessary to tag the dashes (or the dots, but not both) in some way so they could be processed as numbers differing in value from those of the dots.

For the character X the dots go into the dot register as null, null one, one, null; i. e. 6. The dashes go into the dash register as null, one, null, null, one; i. e. 9. This is fully explained on page 16 of the text. The sum of the two numbers is 15, but 15 is a number that could identify at least 11 other characters by that logic. Therefore it *will not* work. Either the dots or the dashes must be tagged.

Tagging an element for this machine is most easily done by multiplying its binary value by 2. By binary value is meant the position of the elements in a step register translated to binary numbers. If the decision has been made to tag the dashes, then for the letter X, the 9 is multiplied by two in an arithmetic unit following inverter **N1**, making a new binary number of 18 for the letter X dashes. Added to the 6 for the dots it makes a new number of 24 for the letter X. The 24 can then be placed in a character register for transmission *directly* to the print wheel where it will match up with its corresponding number as an address on the wheel. Thus the diode matrix and the ASCII generator are eliminated.

Following are the characters identified by the first twenty numbers when developed by doubling the binary value of the dashes.

1 = E	6 = M	11 = D	16 = V
2 = T	7 = S	12 = K	17 = F
3 = I	8 = U	13 = G	18 is blank
4 = A	9 = R	14 = O	19 = L
5 = N	10 = W	15 = H	20 = J

The rest of the Morse code characters can be developed in a like manner. The numeral 9 is 61. The numeral 0 is 62. It should be noted that the first blank number is 18. With 48 numbers being required, 14 numbers will be blank in a six step register. The numbers 0 and 63 cannot be used because of the nature of the Morse code.

In addition to giving a very good clue as to what the diode matrix and the ASCII generator might have been doing they also explain immediately why only five elements of the Morse code are needed to identify a six element special sign. Their last five elements are sufficient for them to fall neatly into one of the *blank numbers. This* would hold true for any combination of six elements so long as their last five elements did not repeat exactly the five elements of one of the Morse code numerals.

Alf A. Jorgenson, KH6AP

tape head cleaners

Dear HR:

In the November, 1971 issue of *ham radio* there is a short write-up by K6KA regarding the cleaning of tape recorder heads. Bill recommends Ampex 087-007 Head Cleaner. Since I've had considerable experience with Ampex machines, both commercial wideband and home entertainment, I though I'd amplify his comments and pass along a warning.

First, most professionals shy away from trichlorethane or carbon tetrachloride, mostly due to their volatile and toxic nature. Alcohol is usually unsatisfactory, since it isn't that good a cleaner and tends to have a drying effect on rubber pinch rollers, causing cracking and flaking of the rubber after extended use.

Ampex Head Cleaner is quite good, but a few precautions are in order. The cleaner is 91% Xylene and 9% trichlorethane. Xylene has one peculiar characteristic: It is an excellent solvent for plastic. Hence, the user must be careful not to spill it on any plastic covers, parts, etc., of the recorder, or they will at least, be marred, and at worst, partially dissolved. In addition, some types of recorders have, in the past, utilized plastic laminate during head construction. Use of Ampex head cleaner on such a head will ruin it in short order! Pure Xylene is far cheaper than the Ampex cleaner, and the only difference is that it is slightly less effective as a cleaner due to the absence of trichlorethane; however, it is perfectly suitable for home applications. Most chemical suppliers can provide it in quart bottles very inexpensively.

Recently, the Navy has dropped most other types of head cleaners in favor of Freon 12 liquid, since it cleans very well, does not attack plastic, vaporizes rapidly, leaves no residue, and is readily available, having already been in wide use for other applications. Commercial cost is unknown, but I suspect that it is fairly low. It can be stored in a sealed glass container at room temperature, and if you spill it, it evaporates in short order, leaving no trace that anything was spilled. It is far less volatile than other cleaners. As a matter of fact, the Navy even uses it for general cleaning; it is available from Federal Stock in spray cans, so there probably is a commercial equivalent.

A word to the wise on cleaning tape recorders: Don't just clean the heads and let it go at that — clean the pinch roller(s), capstan(s), tape guides, etc., and any other surface the tape passes over. If you clean the heads but not the guides, the tape will pick up loose deposits from the guides and transfer them to the heads, requiring more frequent head cleaning. And don't wait until you lose fidelity to clean the heads, or you'll wear them out much too soon. Heads should be cleaned every 10 hours of operation, or, if transcribing records or other tapes, every time the reel is turned over (or its direction of motion reversed). This ensures the best possible recording quality, and long head life.

I have an Ampex 860, purchased in 1967, with 7500 hours of operation and the original heads are just now starting to show a loss of fidelity. That's 312½ days of continuous, non-stop operation! Says a lot for keeping the heads clean. The average home recorder should wear out before the heads go bad, if properly cleaned.

Paul H. Bock, Jr. K4MSG
CWO-2, USN

speech clipping

Dear HR:

I was pleased to get the reference to the Proceedings of the IRE note by Squires and Bedrosian, since it nicely supports my simplified explanation of why speech clipping and the ssb mode are incompatible. (See my article in HR *for February, 1971) I was not aware of the material though I have the Proceedings issue in my collection!*

Let me start by assuring you that rf (or i-f) clipping is indeed well understood.

I believe your difficulty arises from the fact that in the IRE article an infinite bandwidth is assumed. This is why the authors get even worse results for clipped speech in a ssb system than I, because I assumed practical limits for bandwidth (3 kHz) and the lowest audio frequency (400 Hz). An infinite, or to be more practical, a large bandwidth, is a purely relative term and must be viewed against the signal frequency. A 3-kHz bandwidth is indeed nearly infinite when you consider the distortion products or harmonics of a clipped 300-Hz tone. However, a 3-kHz or even a 30-kHz bandwidth is small when you consider a 100-kHz (or 9-MHz) clipped signal.

Since the distortion products or harmonics (multiples of the frequency of the clipped signal) are sufficiently filtered out by a single i-f transformer after the i-f clipper, the problem which is causing your concern does not arise. The subsequent mixers and amplifiers are dealing with an amplitude-limited signal without *the distortion components.*

Some time ago, there was a widespread belief that when you filtered a clipped signal so that all harmonics were removed, you got back the unclipped original. This of course is a fallacy; inspection of the Fourier series for a square wave shows that the output variation is 2 dB for inputs between 1 (the clipping level) and infinity!

Within the context of your letter, clipping at audio is a special case, since the lower harmonics generated by the clipper fall into the band of interest. Clipping at i-f or rf avoids this through the selectivity inherent in most designs so that there is no phase information to lose. (I agree of course that later mixers are in effect ssb mixers.) For example, when clipping a 300-Hz tone, the distortion products are at 900, 1500, 2100, 2700 Hz etc. When you clip a 100-kHz ssb signal, the distortion terms show up at 300 kHz, 500 kHz etc. Obviously, it is no major task to get rid of these, as the band of interest is still only 3 kHz wide.

Walter Schreuer, K1YZW
Ipswich, Massachusetts

freon dangers

Dear HR:

In the July "Comments" column W5PGG said his experience indicates Freon is highly toxic when exposed to high temperatures because it may form phosgene under such circumstances.

I would also like to pass along a warning about Freon, but for a different reason. In my work as a Chapter Director for the American Heart Association, I frequently see research newsletters on investigations sponsored by AHA. The results of a recent investigation of sudden death in kids seeking kicks by breathing aerosol spray fumes, or glue fumes, indicated that they were probably due to heart arrhythmias, that is, abnormal rhythms, such as irregular beat, fibrilation, or complete standstill. In the test, laboratory animals were exposed to strong fields of Freon while their hearts were monitored. The results were those indicated above.

Since Freon is used almost universally as a propellant in every type of aerosol spray, caution is advised. That goes for your wife's hairspray or your tuner cleaner and lube. If it's an aerosol, be sure to use it only in well ventilated areas. The same warning applies to certain types of glue because some of them also contain Freon.

George Baker, WA5RTB
Galveston, Texas

toroid inductors

Dear HR:

We have read with pleasure W6GXN's interesting and informative article in the April, 1971 issue of *ham radio*. I'm sure this will be a great help to anyone building inductive elements.

I notice the author mentioned our old, "Engineers Aide Handbook," which is long since out of print. Its successor (which is soon to be revised), the "Q Curve Book," is priced at a non-inflationary $2.50, postpaid.

R. H. Barden
General Manager, Micrometals

the ham notebook

curtain antenna

While planning to erect a 3-element vertical curtain array I found the only available horizontal supporting wire lacked seven feet in height to raise the elements the required distance above the ground. To overcome the problem, I used light bamboo poles to gain the additional element height above the supporting wire. I used number-18 Copperweld throughout to minimize the overall weight of the array. To duplicate my arrangement, select straight bamboo poles and wind the wire in a long spiral around the poles, providing uniform weight loading to keep the poles straight. Secure the wire elements to the poles with a few turns of monofilament fishing line every 9 inches. For protection against the weather, coat the bamboo poles with two coats of marine spar varnish.

Simple rectangles of 5/16-inch sheet plastic shown in **fig. 2** insulate and support the 3 elements from the horizontal wire. Light weights at the bottom of the outside (director) elements may be used to maintain the bamboo poles erect. Use only sufficient weight to achieve this result, and at the same time avoid unnecessary loading of the horizontal support wire. Number 18 Copperweld wire is rated to support 170 pounds. If the location permits, simply insulate the elements at the bottom and anchor them to stakes driven into the ground.

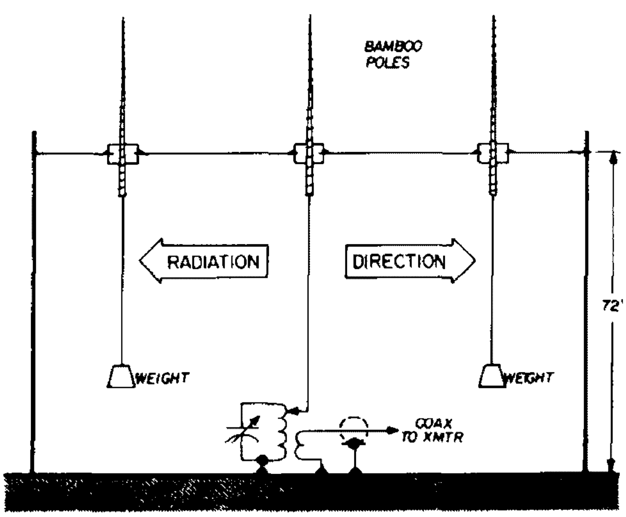


fig. 1. Overview of the "bamboo curtain." The driven element is 68.9', the directors are 65.6' and the elements are spaced 131.2' apart.

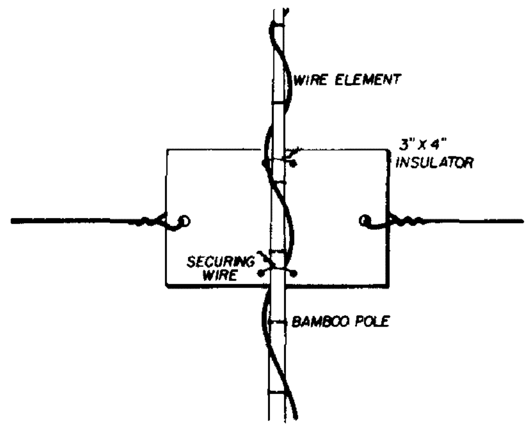


fig. 2. Element supporting insulator details.

The array described is bi-directional in the plane of the elements. The middle element is driven from a tap on the tank, which is resonated at the operating frequency. The tank elements should be of the same value and voltage rating as those in the driving transmitter. The dimensions shown were given for 3.5- and 7-MHz operation. Those interested in operation on 7 and 14 MHz may scale the array to one half the dimensions shown.

Gene Brizendine, W4ATE

swr bridge

The circuit in **fig. 3** is a composite of two articles which have appeared in *ham radio* ("Integrated SWR Bridge and Power Meter," May 1970; "SWR Bridge," October 1971).

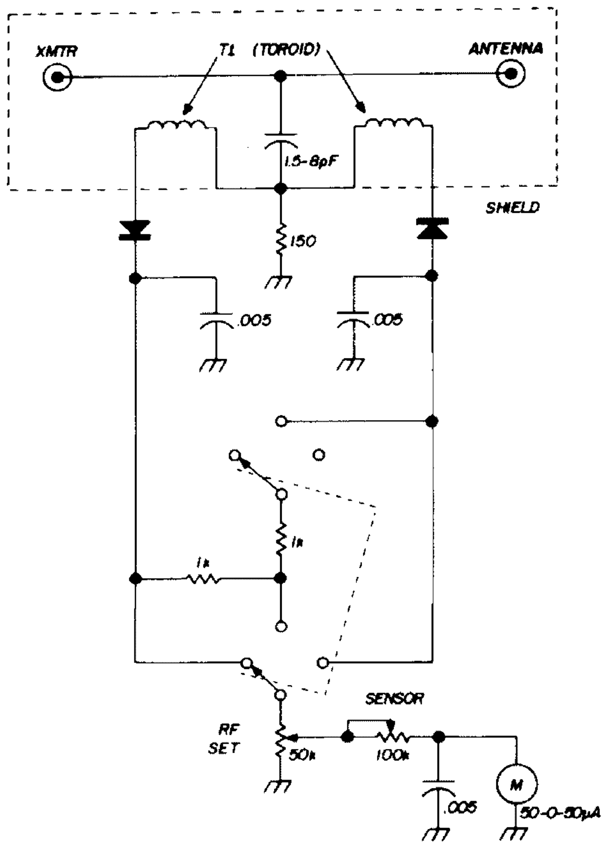


fig. 3. WA5TFK's swr bridge uses a toroid pickup.

I find it rather astonishing that W6DOB's toroid pickup has not received more attention. It takes up very little room, is easy to adjust for a null and is much simpler to construct than the "Monimatch" pickup line.

Output from the toroid at 14 MHz and above was too high for the 100k sensitivity control to handle full scale deflection of the 50-μA meter movement, so a 50k potentiometer was used to make corrections on different bands.

The entire circuit can be constructed in a minibox, size limited only by the size of the components.

Jim Willis, WA5TFK

soldering aluminum

I am surprised to find many amateurs who still do not know how to solder aluminum with ordinary soldering arrangements. This method is certainly not of my own origin, and I pass it on for those who still might be in the dark.

The procedure requires a good, clean, hot soldering iron, a small amount of light machine oil, a pocket knife and some rosin-core solder. Aluminum takes heat like copper, so a big enough iron is essential. *Do not use a soldering gun.*

Place the oil on the aluminum and gently scrape the oil-covered area with the knife until you are sure that the surface is completely cleaned. *Do not remove the oil.* Make sure the soldering iron is clean and well tinned. Rub the cleaned surface of the aluminum with the flat face of the soldering iron while applying pressure. Feed solder to the tip of the iron and the cleaned aluminum will tin just like copper. Once the tinning is complete, other metals or wires which have been tinned can be joined in the usual manner.

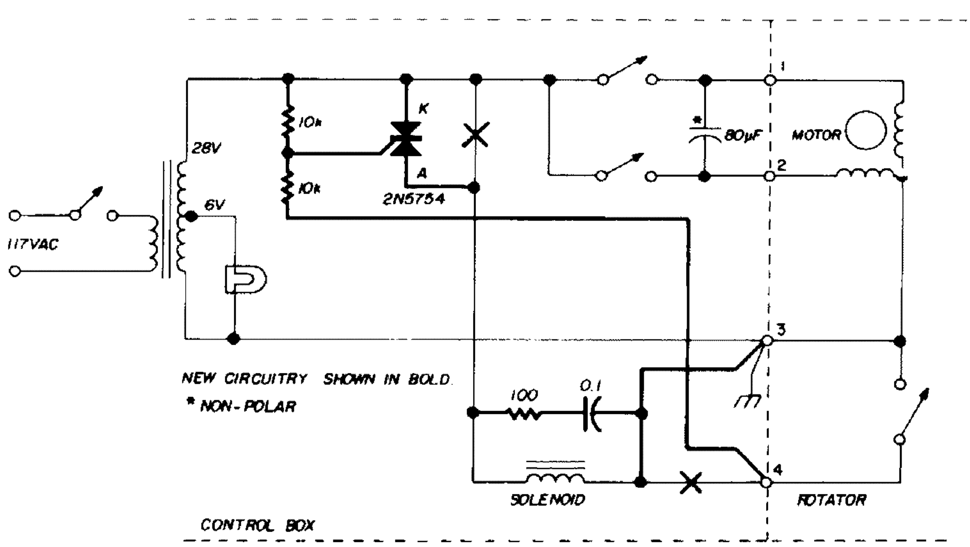


fig. 4. Rotator wiring changes are shown with the heavier lines.

When you first try this technique, do not try either too large surface area or too thick a sheet of aluminum because the iron won't be able to supply enough heat to melt the solder. To prove that the metal is hot enough, place a small piece of solder on the aluminum surface near the tip of the iron but not actually touching the iron. The aluminum must get hot enough to melt the solder for the procedure to be successful. This, of course, applies to all soldering techniques because soldering is only successful when the temperature of both metals to be joined is hot enough to melt solder.

Connections soldered to aluminum, incidently, are vulnerable to attack by moisture which can eventually destroy the joint. Especially important on antennas, the joint should be coated or painted. Just use common sense.

Antenna joints more than five years old are still perfect after being teated with bees wax and PVC tape to keep the moisture out.

Harry L. Booth, ZE6JP

rotator improvement

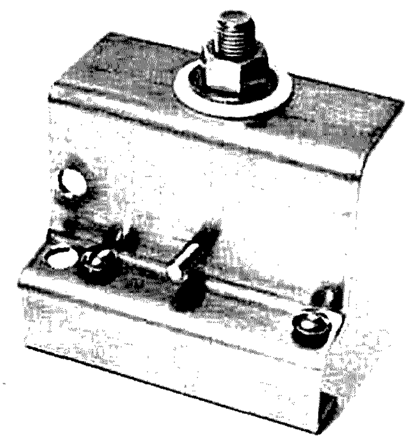
My Alliance model HIR2 rotator (vintage unknown but at least 13 years old) generated such interference on the hf bands that beaming in on a signal of unknown direction was very difficult. The actuation of the solenoid every ten degrees of rotation produced broadband energy which drove the receiver agc quite hard so that attempts to rotate for maximum signal strength were defeated. The problem was cured by adding the simple circuitry shown thus reducing the current in the leads up to the rotator.

The R-C circuit across the solenoid damps transients and protects the triac. The triac conducts the current required to operate the solenoid, confining it to the control box, and only the triac gate current need flow in the leads (3 and 4) to the rotator. Of course, the rotator motor is still supplied via leads 1, 2 and 3, but this steady ac produces no interference.

Les Hamilton, K6JVE/3

new products

no-holes antenna mount



Developed by a ham for his own use, the mobile antenna gutter mount offered by Rejsa Engineering offers a "no holes" method of mobile antenna installation for the hf amateur. The antenna mount is made of very rigid metal and comes complete with all necessary hardware and clear instructions. You can use the mount with a small spring and resonator (not supplied) if desired.

The mount seems to be well engineered and is advertised for amateur, CB, business and police applications. It appears that it would be adaptable to vhf applications also.

The unit is available for $7.95. Information is available by using *check-off* on page 126 or by writing to Rejsa Engineering Company, 7632 Plymouth Avenue North, Minneapolis, Minnesota 55427.

transistor manual

Anyone who works with transistors is familiar with the difficulties involved in locating electrical and physical data for a specific type. Unless the manufacturer is known and his published information is available, a long and possibly fruitless search can result. "Transistor Specifications Manual" has been compiled in order to alleviate these problems. It lists the electrical and physical parameters, along with the manufacturers of nearly *ten thousand* transistor types.

For each bipolar transistor in this manual there is given the polarity (npn or pnp), maximum applied voltages, power dissipation, collector current, operating frequency, collector cutoff current and d-c current gain. A new, separate listing of rf-power transistors includes design frequency, power output, power gain, and collector efficiency in addition to most of the other information. All EIA-registered TO outlines are shown. Where a nonstandard case is used, a dimensioned drawing is provided in a separate section. Transistors are keyed to diagrams that indicate the physical position of the emitter, collector and base terminals. If there is an internal connection to the case, this is noted.

A large quantity of information has been condensed into this "Transistor Specifications Manual" so you can quickly find what you need to know about nearly any transistor.

The new, fifth edition of this handy book is 160 pages long, softbound, and is available from Comtec Books, Greenville, New Hampshire 03048 for $4.50.

More Details? CHECK—OFF Page 126

antenna matcher

The KW107 *Supermatch* is a single package containing a dummy load, antenna switch, swr meter, power meter and the *KW E-Zee Match* antenna tuner. Modern day transceivers must be operated into low-swr terminations for maximum efficiency and long tube and component life. The *Supermatch* was designed to provide a professional uncluttered wife-approved solution to the problem.

The antenna tuner is designed to operate on all bands 80 through 10 meters. It can handle transmitters with power inputs of 1-kW PEP when the natural swr on the transmission line is less than 2:1. For high-impedance feeds the power capability is 350W PEP. The unit works with coaxial and balanced feedlines.

The power dissipation of the air cooled 52-ohm dummy load is a function of time, and is limited by the maximum allowable skin temperature of the glass tinoxide resistor. A graph of time vs average power is included in the instruction book, and ranges from 15 seconds at 2 kW PEP input to continuous service at 90W PEP input.

The wattmeter has two ranges (0-100W and 0-1000W) and measures average power. It is factory calibrated to an accuracy of 5% of full scale. The swr meter in the *Supermatch* is particularly useful because of its high sensitivity. Full scale calibration can usually be achieved with as little as 12W on 80 meters.

The ceramic switch wafers are operated well within their ratings when the other limitations are observed. The KW107 can be arranged for many switching configurations between a dummy load, two 10-, 15- and 20-meter antennas, two 40- and 80-meter antennas and the built-in antenna tuner.

The KW107 *Supermatch* is priced at $134.95 from KW Electronics, 10 Peru Street, Plattsburg, New York 12901. In Canada—KW Electronics, 222 Newkirk Road, Richmond Hill, Ontario. More information is available from these addresses or from *check-off* on page 126.

vhf/uhf mobile antennas

Gain antennas, formerly sold only in the commercial two-way field, are now available for both the 2-meter and 432-MHz amateur bands from Larsen Electronics. The 144-MHz models exhibit 3 dB gain over a quarter-wave whip. The uhf Larsen antennas show 5 dB gain over the same standard.

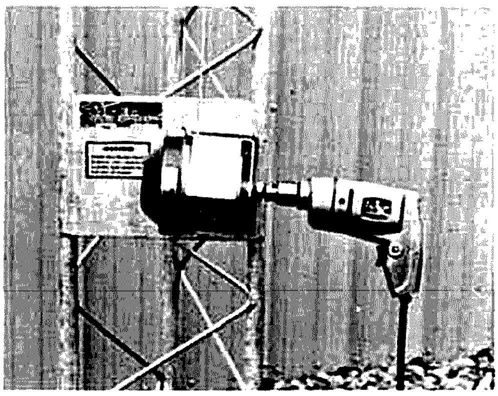
All Larsen antennas feature silver plated stainless steel whips for maximum efficiency. Bases and loading coils are made from hi-impact epoxy with low silhouette design for minimum wind drag. All will handle a full 100W continuous and when installed with sufficient ground plane will have an swr of 1.3 to 1 or less.

The antennas require no coil tuning or special mounting adapters. Complete cutting charts and mounting hardware are supplied. All antennas are fully guaranteed.

The complete 2 meter Larsen Antenna is designated the NLA-150-K. The uhf model is the NLA-440-K. Each comes with all installation materials including coax, plug, snap-in mount and the complete antenna. The price for either is $29. The antennas are also available without coax, mounting hardware and fittings at $24.50. Models can also be obtained to fit all popular types of mount where you supply the mounting hardware.

Complete details are available from Larsen Dealers, by writing directly to Larsen Electronics, P. O. Box 1686, Vancouver, Washington 98663, or by using *check-off* on page 126.

electric winches



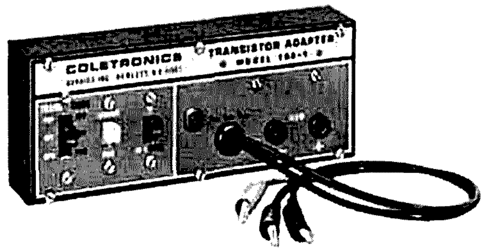
Tri-Ex Tower Corporation has introduced two new electric winches for crank-up towers. The TDD-100 Winch is driven by the average 3/8-inch drill. Two drive bits are furnished with the winch to be inserted into the gear train for raising and lowering. If the drill is not reversible, the short drive bit can be inserted in the opposite end of the winch to lower the load. Braking is automatic and the TDD-100, stopped at any point, will hold a load indefinitely.

The 12 volt electric reversible winch is sold with power cables and an optional battery in weather-tight case and 117 Vac battery charger. It has forward and reverse speeds for raising and lowering the tower, and a level-wind assembly to keep the cable from stacking. Braking is immediate, without coasting or creeping. The electric winch adds a safety feature in eliminating spinning handles, slipping clutches or exposed gears.

Both winches are easily installed and bolt directly onto the existing tower winch mounting frame. Gear train and bearings are packed in lubricant and sealed for life, requiring no additional grease or maintenance.

For complete information on the new electric winches, write Tri-Ex Tower Corporation, 7182 Rasmussen Avenue, Visalia, California 93277, or use *check-off* on page 126.

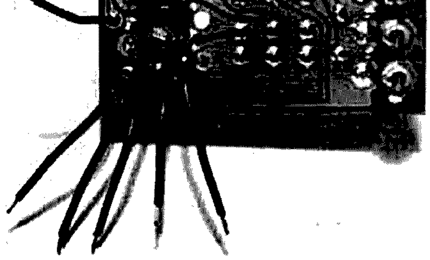
transistor tester



Coletronics has introduced an adapter to convert any vom to a transistor and diode tester for out-of-circuit devices. The adapter tests both high-and low-power npn and pnp transistors for shorts, leakage, open circuits and current gain. Diodes can be tested for opens, shorts, and both forward and reverse current without transferring test leads.

The unit comes with instructions and two batteries. It sells for $14.95 plus $1 shipping. More information is available from the manufacturer, Coletronics Service Inc., 1744 Rockaway Avenue, Hewlett, New York 11557 or use *check-off* on page 126.

regency crystal board



Topeka FM Engineering is selling a new, printed-circuit crystal deck for add-

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ing six more transmitter frequencies to the Regency HR-2, HR-2A and the Transcan. The 2½ x 1½-inch board has plug-in crystal sockets, fixed NPO capacitors, Teflon wiring and high-quality ceramic trimming capacitors. It comes with all necessary mounting hardware and mounting instructions. The boards are available in kit form for $9.50 and wired for $13.50. Shipping for either board is 60c.

More information is available by using *check-off* on page 126 or by writing to Topeka FM Engineering, 1313 East 18th Terrace, Topeka, Kansas 66607.

antenna coupler

Designed to complement the popular 100- to 200-watt ssb transceiver, a new antenna coupler by RF Communications can match the nominal transceiver 50-ohm output to a resonant doublet or any wire at least 15 feet long. The RF-675 coupler covers the frequency range of 2 to 30 MHz and guarantees a vswr — as seen by the transceiver — of 1.2:1 or less. The unit, which can handle 100-watts average power or 200-watts PEP, has a built-in dummy load for off-the-air tune-up and digital turns counting for rapid resetting of the tuning controls. There is also metering of output, forward and reflected power.

For more information write to RF Communications, Inc., 1680 University Avenue, Rochester, New York 14610 or use *check-off* on page 126.

antenna accessories

Apollo Products offers a vary complete line of medium-power amplifiers, wattmeters, antenna switches and cabinets. Included in this line is the model 700X rf wattmeter and dummy load covering 80 through 2 meters and featuring power ranges to 10, 300 and 1000 watts along with modulation percentage on transmitters with less than 100 watts output.

The unit offers meter accuracy of 5% below 60 MHz and 8% above 60 MHz. It is completely portable and very ruggedly built. The unit, a passive instrument, comes with a built-in handle, lending itself to portable and field usage. For use, simply connect the transmitter and antenna into two standard SO-239 chassis connectors.

The unit sells for $124.50. More information is available from Apollo Products, Box 245, Vaughnsville, Ohio 45893 or by using *check-off* on page 126. The Apollo Products catalog includes the complete line of equipment. Apollo also produces a series of deluxe cabinets for home construction. A flyer describing them is available on request from the same address.

transistor design book

In the second edition of *Practical Design with Transistors,* Mannie Horowitz provides the kind of information that will allow a serious experimenter, engineer or technician to design a transistor circuit from scratch. Much of the cut-and-try work is eliminated by use of the manufacturers specification sheets and appropriate formulas.

Mr. Horowitz provides practical design data and equations but does not devote many pages to equation derivations or the solid-state physics behind transistor operation. The book is overtly a guide to efficient, practical circuit design—not an engineering or physics textbook. The publisher claims that anyone with a working knowledge of algebra and radio elec-

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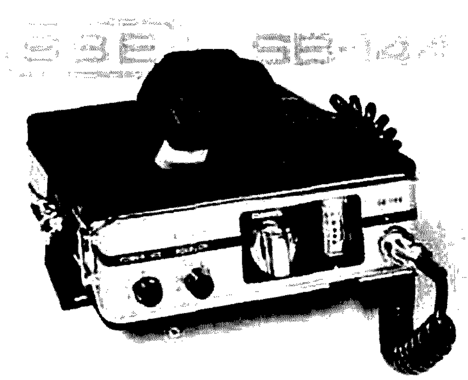
tronics should have no difficulty in designing a transistor circuit with this book.

In order to give the reader a better understanding of design procedure, the characteristics and problems of each particular type of component or circuit configuration are discussed. This is followed by a description of the equivalent circuits and the mathematical relationships of the various parameters. Examples are presented throughout the text to demonstrate the practical use of the final circuit equation as it is applied to determine specific circuit components.

Special consideration has been given to bias stabilization, frequency response, coupling, phase inversion and feedback. Amplifiers, multivibrators and voltage regulators are covered thoroughly as are the characteristics and design applications of silicon controlled rectifiers, unijunction transistors and fets.

The book is filled with clear schematics, diagrams and curves to provide a valuable information source on transistor circuits for the serious experimenter or engineer. This Howard W. Sams book is 288 pages, softbound and costs $6.95. It is available from Comtec Books, Greenville, New Hampshire 03048.

fm transceiver



Linear Systems has introduced a new two-meter fm transceiver, the SBE model SB-144, to complement the new SBE-36 high-frequency ssb transceiver.

The new fm transceiver features 10 watts output, 12 channel capability and a built-in S-meter/rf output indicator. The unit comes with a dynamic microphone and crystals for the three most popular repeater frequency combinations presently in nationwide use.

More information is available from Linear Systems, Inc., 220 Airport Boulevard, Watsonville, California 95076 or by using *check-off* on page 126.

10-MHz oscilloscope

Heathkit is marketing the IO-103, a new 5-inch oscilloscope kit with triggered sweep and response from dc to 10 MHz. The scope features a 6 x 10-cm screen, less than 50-ns vertical rise time for a 50-ns/cm sweep rate, triggered sweep with selection of either normal or automatic modes, switch controlled ac-dc coupling, provision for external triggering signals and horizontal deflection signal, dual 120/240 Vac power supply and front panel connectors for vertical inputs and 1V peak-to-peak signal for checking calibration. The unit costs $229.95.

Full details on the oscilloscope are available in the latest Heathkit catalog which is available free by using *check-off* on page 126 or writing to Heath Company, Department 122-2, Benton Harbor, Michigan 49022.

automotive electronics

The active experimenter frequently finds himself working with automotive electronics—installing mobile rigs, theft alarms, modern ignition systems and electronic accessories, along with servic-

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ing the increasingly complex family car. Far from the simplicity of the Model T, today's car features continually evolving starting, charging, ignition, lighting and indicating systems. Also, more and more electronic test equipment is coming out for car repair—CRT engine analyzers, dwell meters, tachometers and timing gear.

"Automotive Electronics" by Rudolf Graf and George Whalen is 320 pages of text and illustrations exploring the electronics in the modern automobile. Although not directly intended for the radio amateur or experimenter, it is a fascinating book in its own right and also can serve as a handbook of the components, circuitry and test gear common to the ham shack and the garage.

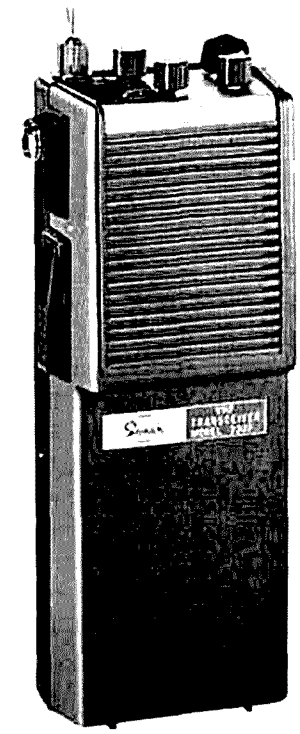
The new book, published by Howard Sams, is softbound, 320 pages and costs $6.95 from Comtec Books, Greenville, New Hampshire 03048.

pushbutton transceiver

The new Sonar model FM 3601 10-watt fm transceiver features pushbutton selection of its eight channels. The rig is all solid-state and includes netting trimmers for all receive and transmit crystals on the military-grade, glass-epoxy printed-circuit board. Serviceability, compactness and receiver overload protection are built in.

The unit sells for $299.95, including a microphone and crystals for 146.94 MHz simplex and 146.34/146.94 MHz repeater use. A descriptive brochure including all the impressive specifications of this new transceiver is available by using *check-off* on page 126 or by writing to Sonar Radio Corporation, 73 Wortman Avenue, Brooklyn, New York 11207.

hand-held transceiver



Sonar has introduced a five-channel two-meter hand-held transceiver with a minimum 1.6W output. Weighing only 1½ pounds without a snap-in rechargeable battery option, the hand-held unit can run for 8 to 14 hours on a battery charge. Standby current drain is only 10 mA. The unit also boasts FCC type acceptance for use in non-amateur services. Circuitry includes ceramic and crystal band-pass filters and transmitter protection against antenna mismatch. The unit uses all electronic switching (no relays) and military grade components and pc boards throughout. There are connections for optional external antenna, earphone and battery charger. A wide range of accessories and power supplies are offered.

The basic unit sells for $450.00 with collapsible antenna and shoulder strap, less crystals and nicad battery cartridge. The rechargeable nicad pack is $30.

The complete specifications on this flexible unit complete with a list of options is available by using *check-off* on page 126 or by writing to Sonar Radio Corporation, 73 Wortman Avenue, Brooklyn, New York 11207.

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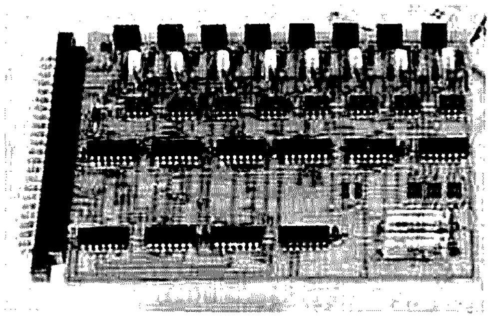
tone decoder



A flexible new tone decoder is available from Lee-Com Associates of Cleveland, Ohio. While lending itself primarily to vhf repeater control, the unit is also suitable for data communications terminals, traffic control, station control and security systems.

The Multone decoder can decode 12 functions (3-by-4 pad) with high-true TTL outputs in decimal form and high-true outputs in 4-line BCD (0-9). It can decode 7 tones from a standard pad with 100 mA low-true sink drive for bulbs and relays. It has separate high and low groups HIT out for single tone detection and two-tone PARITY output for simultaneous tone detection.

The unit operates from a 5-Vdc power supply and is built on a military grade glass-epoxy pc board. The board measures 4½ x 6½ inches. Options include light-emitting diode lamps to indicate 7 or 8 demodulated tones, and a version of the decoder to handle 16 functions, 8 tones (4-by-4 pad). The basic decoder sells for $169.95, the 16 function decoder for $189.95 and the 8 LED display sells for $18.00.

Complete information is available from Lee-Com Associates, Box 43204, Cleveland, Ohio 44143 or by using *check-off* on page 126.

heath digital frequency display

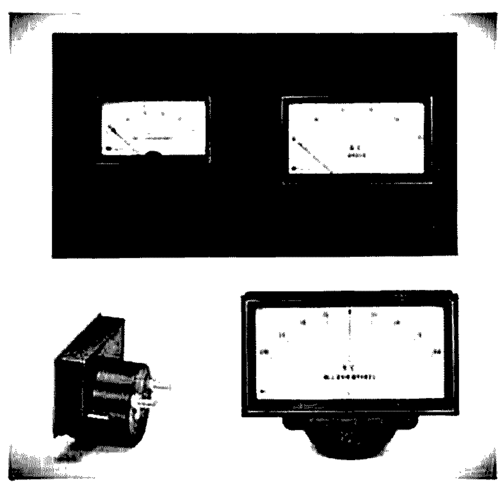
The Heath Company has introduced the SB-650 Digital Frequency Display, a digital computer for use with Heathkit receivers and transceivers that calculates operating frequency and displays it via numerical readout tubes as the amateur tunes across the band. The unit is usable with all Heathkit SB-series and HW-series multiband transceivers.

Unlike any other counter on the market, the SB-650 actually computes a received or transmitted frequency using the three different frequencies produced by the heterodyne oscillator, LMO and bfo in the receiver/transceiver. Six readout tubes display the frequency in MHz, kHz and hundreds of Hz with accuracy within 100 Hz ±1 count.

Priced at $179.95 mail order, the Heathkit Digital Frequency Display boasts all solid-state circuitry with 35 ICs and six transistors. The only control is an external on/off switch. Four internal adjustments permit reducing generated spurious frequencies to lower than 0.25 μV equivalent signal level.

More information on this new display is available by using *check-off* on page 126 or by writing to the Heath Company, Benton Harbor, Michigan 49022.

meters



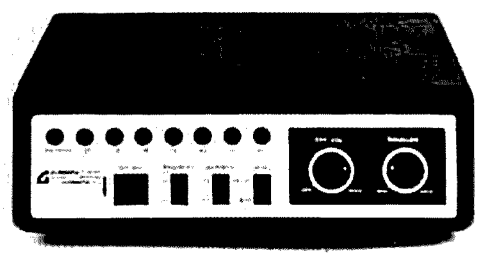
A new designer-styled *GL/B* panel meter series has been introduced by Triplett. Flat and rectangular shaped, the meters feature glass windows with a modern front resembling a bezel. Similar to bezel mounted meters, the new meters mount easily behind the panel but they do not need the normal bezel hardware, thus saving cost and installation time. The new meters take up a minimum of panel space and their raised border can be painted or printed to further enhance a panel.

Two-stud mounting is standard on all of the new *GL/B* series panel instruments. Three basic sizes are available, 3½, 4½ and 5½ inches. The case barrel of the new *GL/B* series, housing the meter movement, has been designed so that it extends only 1.840 inch from the panel mounting, saving valuable back-of-the-panel space and permitting quick mounting. Back-of-the-panel supporting brackets are provided with each new Triplett panel instrument in the series.

Both pivot and jewel movements, as well as suspension types, are used in the new panel instruments. They are available in many ranges of dc voltmeters (1000 ohms per volt), dc and ac milliammeters, dc and ac ammeters, dc microammeters, dc millivoltmeters, rf thermoammeters, decibel meters, ac voltmeters (rectifier type) and volume unit meters.

The new *GL/B* series, with the standard two-stud mounting, is available from Triplett Modification Centers and company stocks. For additional data, write to Triplett Corporation, GL/B Department, Bluffton, Ohio 45817 or use *check-off* on page 126.

scanning receiver



Gladding has introduced a new vhf scanning receiver with a number of unusual features. Off-the-shelf, it covers not just the two-meter amateur band but up to eight crystal-controlled frequencies anywhere from 144 to 175 MHz. The unit also offers a priority channel to which the set will automatically revert regardless of reception on other channels. The HiScan has a built-in 117 Vac and 12 Vdc power supply and a trap door for rapid crystal changing. The unit sells for $114.95.

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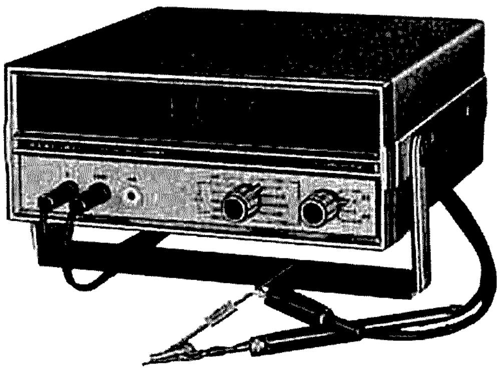
More information is available by using *check-off* on page 126 or by writing to Gladding Corporation, Pearce-Simpson Division, Box 800 Biscayne Annex, Miami, Florida 33152.

digital multimeter



Heathkit is offering a new digital multimeter boasting 0.2% accuracy, a built-in calibrator and 3½ digit readout. A "half-digit" readout tube indicates only the digit one rather than 0-9. It can be used to change a meter with a maximum readout of 999 to one reading 1999.

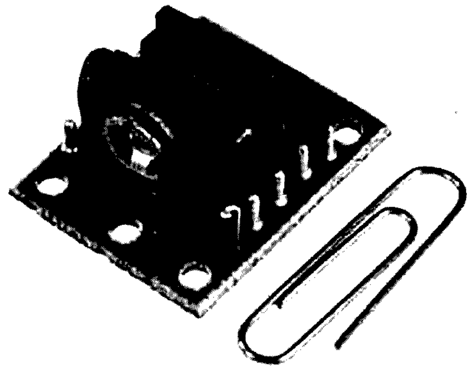
Features of the new meter include the ability to change polarity measurements without having to change leads or throw a switch. The meter reads ac voltages from 100 μV to 500V, dc from 100 μV to 1000V, current from 100 nA to 2A and resistance from 0.1 ohms to 20 megohms. Input impedance is 1000 megohms on the 2V range, 10 megohms or higher on the other ranges.

The decimal point is placed automatically and the meter has built-in overload protection and a front-panel overrange indicator.

A dc calibrator is furnished and an internal circuit and unique transfer method allow accurate ac calibration. Solid-state design with cold-cathode readout tubes and memory circuit give stable, non-blinking operation. The kit includes standard banana jack connectors complete with test leads. The kit sells for $229.95, plus shipping.

More information is available in the latest Heathkit catalog, available by using *check-off* on page 126 or by writing to the Heath Company, Department 122-2, Benton Harbor, Michigan 49022.

miniature tone encoder



Alpha Electronic Services has introduced a new thick-film hybrid tone encoder model ST-85J, designed for use in portable and hand held communications equipment. This compact, easily-installed unit provides an economical method of controlling repeaters, remote base stations or special functions. When used where CTCSS tone quieting is desired, the ST-85J will provide the necessary sub-audible tone to activate the decoder in the base station. For special control functions, paging, or selective repeater entry, the unit is available in high frequencies as well as sub-audible (20 Hz to 3000 Hz).

The thick-film hybrid technique along with the use of ICs achieves extremely small size and long-term reliability. Frequency stability is ±0.5% over a temperature range of -30° to +100° C and current requirements are less than 4 mA at 12.6V.

The unit comes with a number of different installation kits and step-by-step instructions for easy installation in any make or model radio. The unit sells for $47.00, postpaid. More information is available by using *check-off* on page 126 or by writing to Alpha Electronic Services, Inc., 8431 Monroe Avenue, Stanton, California 90680.

plugbords

Vector has introduced a new line of miniaturized plugbords specifically designed for mounting dual in-line packages and discrete components with leads spaced on 0.1-inch increments. All of the plugbords have an overall grid of 0.042-

More Details? CHECK–OFF Page 126

inch holes on 0.1-inch centers and are 2.7-inches wide by 4.5-inches long in economical phenolic or mil-spec epoxy glass material with a choice of one or two ounce copper etched contacts on 0.156-inch or 0.1-inch centers respectively. Contacts are tinned or nickle-plated and gold-flashed for high reliability applications.

To insure widest possible usage, a variety of etched patterns are offered on the boards. Patterns offered include models with interleaved vertical buses for Vcc and ground and groups of three hole pads for IC mounting, or models with buses only, which are intended for mounting sockets for wire wrapping. For users of hybrid circuitry, models are available without any pattern on the upper portion of the board, and Vector Edge Pin contacts which permit nearly any style hookup.

To complete the package, the firm has a complete line of terminals and sockets to go with the boards as well as card cages and extruded cases for packaging the plugboards.

Prices of the boards range from $2.19 to $5.55 and quantity discounts are available. Boards are available now from the firm's authorized industrial distributors. For more information write to Vector Electronics Company, 12460 Gladstone Avenue, Sylmar, California 91342 or use *check-off* on page 126.

miniature microphones

Two styles of miniature push-to-talk microphones for use specifically with 2-way communications equipment are

now being offered by Controlonics Corporation of Littleton Common, Massachusetts.

These new magnetic microphones are designed specifically for use with personal portable or hand-held transceivers. The Mini-Mike is 2-inches long, 5/8-inch wide and 15/32-inch thick and weighs 1/2 ounce with cord. Response is centered on the 300- to 5000-Hz range with a sensitivity of minus 76 dB at 1000 Hz. The dependable magnetic microphone element is self-shielded against magnetic fields and has a nominal impedance of 4000 ohms.

The Micro-Mike is 1-inch long, 0.365-inch wide and 0.285-inch thick and weighs 1/10 ounce. The useful frequency is from 400 to 5000 Hz and sensitivity is minus 80 dB at 1000 Hz.

More information is available from Controlonics Corporation, One Adams Street, Littleton Common, Massachusetts 01460 or by using *check-off* on page 126.

alignment tools

Jensen Tools offers a new kit of 25 tools for the advanced amateur, experimenter and serious electronic technician. Designated 23C750, the kit includes virtually every alignment tool needed for work on radio-frequency electronic circuits used in amateur, mobile, marine and TV equipment.

Each tool provides the necessary isolation between operator's hand and the equipment tested. Included in the complement of tools are a universal aligner, long-reach core aligner, extra-thin tuning wand, bone fiber tuner, Delrin-tipped i-f transformer aligner, oscillator aligners, and special TV aligners (for Motorola, Stewart-Warner, Belmont, Zenith, RCA and Westinghouse). Working ends include slotted, recessed and hex styles, ranging in tip size from 1/32-inch to 1/4-inch.

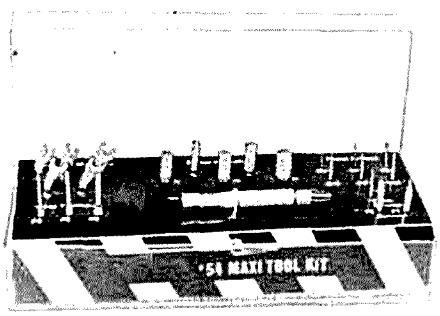
The 25 precision tools are furnished in a rugged roll pouch which fits conveniently into tool chest or desk drawer. The pouch is designed for easy removal and replacement of tools, and has a fold-over

 More Details? CHECK—OFF Page 126

flap which prevents tools from being lost.

The 23C750 alignment set is available for under $17.00, postpaid. For further information contact Jensen Tools and Alloys, 4117 North 44th Street, Phoenix, Arizona 85018 or use *check-off* on page 126.

subminiature tool kit



Tiny multi-purpose instrument tools in a 22-piece set are now available for use on very small fasteners. Suitable for use on meters, clocks, models and miniaturized projects, the tools are designed for instrument repair and light assembly work involving very small nuts, set screws and machine screws.

The tool set, Mini Kit Number 54, contains six jewelers' type screwdrivers in sizes 0.025, 0.040, 0.050, 0.070, 0.080 and 0.100 inch, two cross-recessed Phillips-type drivers in sizes 0 and 1, five open-end wrenches in sizes 5/16, 3/32, 7/64, 1/8 and 5/32 inch, three Allen-hex type wrenches in sizes 4, 6 and 8, five socket wrenches sizes 5/64, 3/32, 7/64, 1/8 and 5/32 inch, a marking scribe and a knurled chuck-type handle for positive gripping. The blades are interchangeable, and all are made of hardened, tempered nickel-plated tool steel.

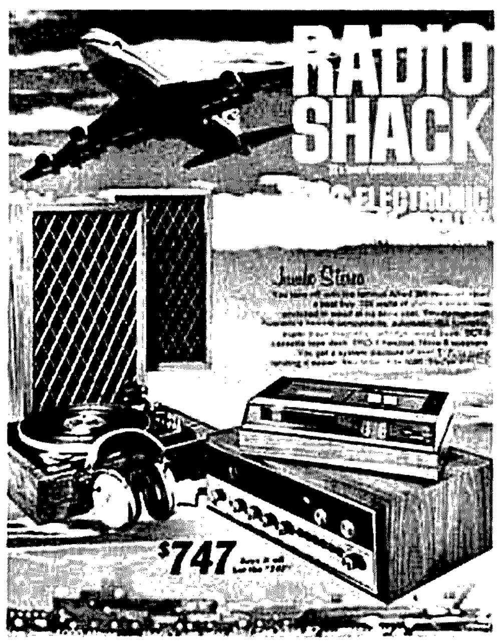
The complete set is packaged in a clear plastic box for easy use and convenient storage. It is priced at $14.50, postage paid. For further information, contact Jensen Tools and Alloys, 4117 North 44th Street, Phoenix, Arizona 85018 or use *check-off* on page 126.

strip printer

The details of the strip printer used in the "Automatic Fist Follower" in the November, 1971 issue of *ham radio* were accidently deleted. The unit in the lead photograph is the Computer Terminal System's Model 4 strip printer. This little unit prints at 35 characters per second on pressure-sensitive 15/32-inch paper tape. The entire alphabet, digits 0 through 9 and 28 symbols make up the 64 characters which this machine is capable of printing. While other print codes are optional, ASCII is standard. The unit is normally powered by 117 Vac, but optional 220-Vac and 12-Vdc versions are available.

The units feature modular construction and extensive use of ICs. Many options are offered. It measures 4-1/8 x 8-7/8 x 3-inches. The complete mechanical and electronic specifications on this unit and the entire line of CTS strip printers are impressive. They are available from Computer Terminal Systems, Inc., 52 Newton Plaza, Plainview, New York 11803 or by using *check-off* on page 126.

electronic equipment catalog



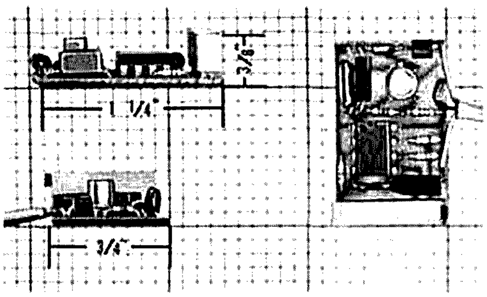
An amateur bands receiver, an SWL receiver and a parcel of vhf monitors and

More Details? CHECK—OFF Page 126

Citizens Band equipment are included in the 1972 Radio Shack Electronic Equipment Catalog. Along with this communications gear, the 92-page full-color catalog includes a wide range of audio and home entertainment equipment, public address systems, tape recording and cassette gear and a selection of intercoms and hi-fi speakers.

The new catalog is available free at any Radio Shack store. Allied Radio Shack, 2617 West Seventh Street, Fort Worth, Texas 76107.

walkie-talkie encoders



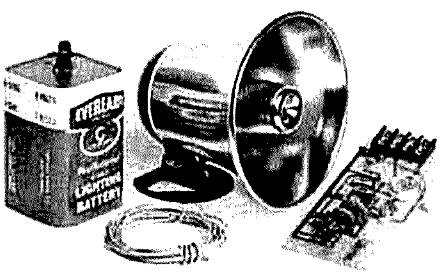
Ross and White have expanded their line of tone encoders for entry into tone access fm repeaters. The most recent encoder is the model TE-P, a single tone, factory-preset tone-burst encoder for installation in hand-held fm transceivers. The unit is field adjustable over the range of 1700 to 2500 Hz. It features all American manufacture, IC design, low current drain and adaptability to almost any portable. The unit is tiny, measuring 1¼ x ¾ x 3/8 inches and comes with mounting details for your specific transceiver. Ross and White will be arranging factory installation of the encoders for those who do not want to handle the installation in the limited spaces in most walkie-talkies. The units uninstalled, sell for $34.95, postpaid.

The complete specifications on the unit are impressive and are all included in an information sheet complete with oscillograms of the audio burst envelope and output waveform. The sheet is available by writing to Ross and White Company, 50 West Dundee Rd., Wheeling, Illinois 60090 or by using *check-off* on page 126.

limited to, young people. Assuming no previous technical understanding by the reader, it thoroughly explains the fascination of amateur radio, the jargon, the gear necessary, how to get a license, learn the code, select gear, erect an antenna, use simple test gear and acquire good operating procedure. All in all it answers any conceivable question that the neophyte might have. There are also two chapters devoted to careers for amateurs in the military, industry and broadcasting.

This book should be in every town library and certainly in every high-school library. It would be handy in the ham shack to lend to inquiring neighbors, newspaper boys and coworkers. If you think there is a likely candidate to become a ham, giving him this book just might do the trick. Softbound, 185 profusely illustrated pages, $5.25 from Comtec Books, Greenville, New Hampshire 03048.

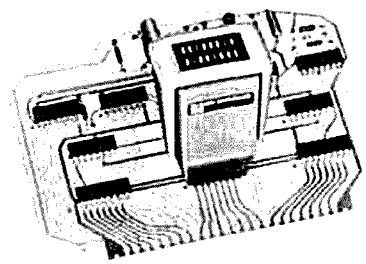
burglar alarm



Intended for many installations, but particularly suitable to protecting mobile rigs, the new ECD siren module offers the heart of an effective car burglar alarm system. The siren module kit contains a pre-assembled module which emits a two-tone wailing audio sound rated at 96 dB at 25 feet (10 watts output) when run on 12 volts. It can also be run on 6 volts, however. Also in the siren kit are a projection speaker, a 6 volt battery, 100 feet of hookup wire and instructions. The system sells for $24.95 plus postage. The pre-assembled module alone costs $14.95.

For more information write to Electronics Circuits and Design Company, Inc., 33 East Chestnut Street, Alliance, Ohio 44601 or use *check-off* on page 126.

logic checker



A small, portable, passive logic checker has been introduced by ALCO Electronic Products. The Model 101 Logic Checker gives instantaneous visual indication of the logic states of all terminals on TTL and DTL 14- and 16-pin dual-in-line integrated circuits.

The 101 is totally self-contained and requires no batteries or power cords—it is powered by the circuit under test. A bank of small incandescent lamps on the top of the unit along with a set of 24 templates covering virtually all popular IC packages display and interpret the logic levels present on all terminals as long as the package uses standard logic voltages between 0.5 and 5 Vdc.

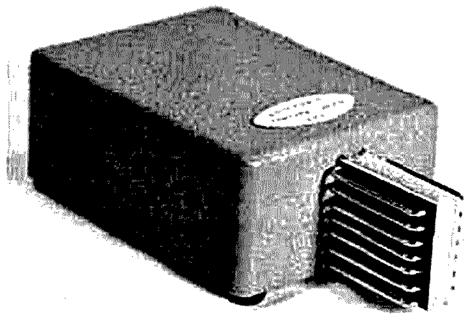
The unit's overall dimensions are 1.7 x 2.65 x 1.35-inches. The complete specification sheet on the unit is impressive and is available from Alco Electronics Inc., Box 1348, Lawrence, Massachusetts 01842 or by using *check-off* on page 126. This handy piece of test equipment sells for $89.95.

introduction to ham radio

How do you explain to a curious youngster what our hobby is all about? How do you explain the difference between CB and amateur radio, or how do you interpret ham jargon? It can take a lot of explaining and it can get pretty confusing to a young person. Robert Hertzberg, W2DJJ, has come out with the fifth edition of, "So you want to be a Ham," and this time-tested manual goes a long way toward telling the prospective amateur all about the hobby.

With hundreds of photographs, this book is written for, but definitely is not

ic electronic key



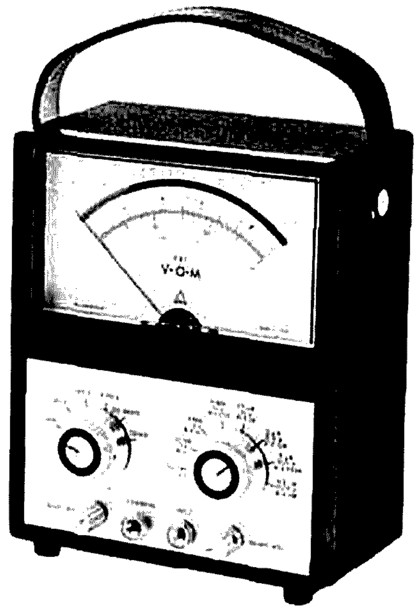
The Electronic Feather Touch Key is a completely solid-state key without paddle movement, spring tension or contact clearance adjustments or any keying contacts. Keying problems caused by contact corrosion and bounce as well as paddle tension and movement adjustments are no longer a concern. The key is the natural interface for the extremely low current, high speed integrated circuits used in solid-state electronic keyers.

The Electronic Feather Touch Key is essentially an electronic single-pole double-throw switch activated by finger touch. Touching the electronic grid paddle causes a silicon monolithic integrated level detector to conduct. Conduction of each level detector causes its output to remain at ground potential as long as a high-impedance short (finger) remains across two adjacent grids. The design of the grid paddle prevents erratic keying due to finger moisture.

The small size key was designed specifically to meet the needs of today's solid-state keyers and to eliminate those problems experienced with mechanical keys.

The key is built on a glass epoxy board housed in an attractive light blue cabinet. It can be powered by internal batteries or 3 to 4.5 Vdc at 10 mA can be taken from a keyer. The unit is weighted to keep it in place during operation. The new key is guaranteed for a year and sells for $19.95 wired and tested. More information is available from Data Engineering, Inc., Box 1245, Springfield, Virginia 22151, or use *check-off* on page 126.

fet vom



The new Delta model 3000 fet vom is billed as the bridge between the multimeter and the digital voltmeter. It features ten-turn adjustment potentiometers for zeroing and ohmmeter adjustment; 10 megohm input impedance, and a variety of meter ranges which will read resistance from zero to a billion ohms, current from one nano-ampere to 300 milliamperes and both ac and dc voltages from 0.3 volts to one kilovolt.

The metering circuit incorporates a lot of innovations for a $74.95 ready-built unit ($59.95 kit). The meter is in the feedback loop of an ic operational amplifier which, among other benefits, minimizes nonlinearity in ac readings. There is current limiting in the ohmmeter circuit to prevent destruction of most semiconductors under test. There is also an R-C filter in the dc metering circuit to minimize inaccuracies due to stray rf pickup. The meter itself is a six-inch D'Arsonval mirror-back movement.

For complete information on these meters – this description just scratches the surface – write to Odyssey Equipment Company, Box 3382, Madison, Wisconsin 53704 or use *check-off* on page 126.

The magnetic feature is being offered on midget pocket clip, regular, extra long, and super long fixed handle drivers and also on interchangeable shanks for use with the Xcelite Series "99" handles, both regular and ratchet types.

An Alnico permanent magnet inserted in the socket holds fasteners firmly for one-hand starting, driving or retrieving hex screws, bolts and nuts, in close quarters and hard-to-reach places. The magnet is insulated so that the tool socket itself remains unmagnetized and will not attract extraneous matter or be deflected by nearby metal surfaces. Nutdriver sockets are specially treated and hardened to withstand severe service, such as the driving of hex head self-tapping screws. They are finished in black oxide for dimensional control as well as for quick identification as magnetic tools.

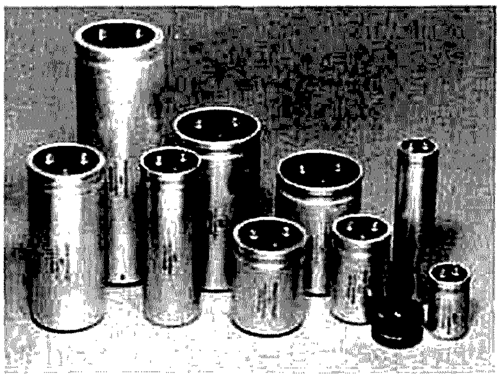
Fixed handle drivers have color-coded plastic handles in a new comfort-contour design. This further distinguishes the magnetic from regular Xcelite nutdrivers and provides adequate torque for one-hand driving. American made, the magnetic nutdrivers are available nationwide through local distributors. Complete information may be obtained by requesting bulletin 671L from Xcelite Incorporated, Orchard Park, New York 14127 or by using *check-off* on page 126.

ic op amp book

Theory, circuitry and applications of the increasingly important integrated-circuit operational amplifier are covered in a new book by Roger Melen and Harry Garland. "Understanding IC Op Amps" contains chapters on the ideal op amp, real-life limitations of actual op amps, basic semiconductor theory, op amp circuitry, practical design considerations and basic op amp terms and definitions. Many practical circuits – complete with component values – are included in this book.

Published by Howard W. Sams. 128 pages, softbound; the book is available for $3.95 from Comtec Books, Greenville, New Hampshire 03048.

computer electrolytics



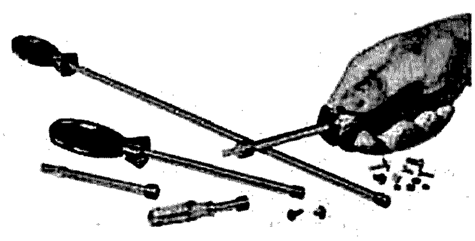
To meet changing requirements for higher-capacitance, smaller-size capacitors in computer power supplies, industrial control equipment and energy-storage applications, a series of 190 extended capacitance ratings has been added to the Sprague Type 36D Powerlytic aluminum electrolytic capacitor line.

Some of the new capacitors feature up to 90% more capacitance in a given case size than was previously available. Type 36D Powerlytic capacitors now have a capacitance range from 50 to 650,000 μF, with voltage ratings from 3 to 450 Vdc.

Other features include increased ripple current capability, low leakage currents, operation at 85°C without derating and extremely long shelf life.

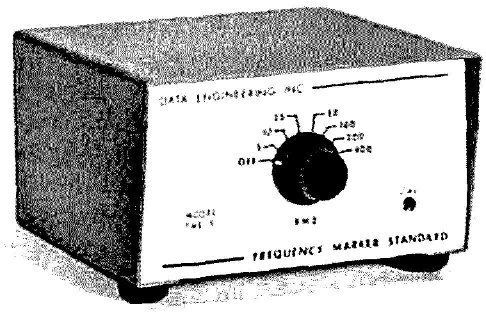
Type 36D Powerlytic capacitors are available for off-the-shelf delivery from Sprague industrial distributors. For futher information write to Sprague Products Company, North Adams, Massachusetts, 01247 or use *check-off* on page 126.

magnetic nutdrivers



Amateur experimenters, along with professional service men may benefit from the addition of 1/4- and 5/16-inch hex socket magnetic nutdrivers to Xcelite's line of professional hand tools.

frequency marker standard



Data Engineering has introduced a frequency marker standard with a choice of seven different marker frequencies: 5, 10, 25, 50, 100, 200 and 400 kHz. Harmonics are usable above 50 MHz, and units have a front-panel adjustment for calibration against WWV. The square-wave output features better than 10-nanosecond rise and fall time and 3.8-volt peak-to-peak output.

The Data Engineering FMS-3 can be used to calibrate receivers, grid dippers, signal generators and oscilloscopes. It can be used in testing amplifiers for damping, ringing, frequency response and bandwidth. The unit also makes an accurate and stable square-wave generator capable of driving the most demanding logic circuits.

The circuit of the FMS-3 includes a precision 400-kHz crystal and a complement of Fairchild ICs. It is designed to give usable outputs above 50 MHz yet eliminate unwanted markers. Units are guaranteed for one year, are built on glass epoxy boards and come in an attractive blue and gray cabinet. The factory assembled and aligned version sells for $32.95. The identical circuit – wired, tested and calibrated but without the cabinet, switch and battery holder is priced at $22.95; the complete kit without the cabinet, switch and battery holder sells for $19.95.

More information is available from Data Engineering Incorporated, Box 1245, Springfield, Virginia 22151 or by using *check-off* on page 126.

More Details? CHECK–OFF Page 126

house wiring guide

Too often amateurs tend to regard house wiring as requiring only common sense and Ohm's law. "House Wiring," the latest in the Audel series of technical books, explodes this myth by explaining the planning and actual installation which produces a safe and functional professional job. Obviously the results of years of practical experience, this book is guided paragraph by paragraph by the National Electrical Code and explains such things as the practical techniques of wiring house additions and mobile homes and adding the necessary wiring to run the highest power linear amplifier or the most well equipped workshop. Fusing, safety standards and load calculations are all included in this handy book. It showed one amateur, at least, that there is a lot more to house wiring than meets the eye.

"House Wiring" by Roland E. Palmquist is 183 pages, hardbound, and costs $5.95. It is available from Comtec Books, Greenville, New Hampshire 03048.

capacitor tester

The model CT-1 Capacitor Tester and Leakage Indicator allows dynamic testing of all types of capacitors for leakage, opens and shorts. The built-in power supply allows the CT-1 to safely reform aged electrolytic capacitors along with providing for dynamic capacitor testing. In addition, the unit can test vacuum tubes for inter-element leakage and open filaments. It also serves as a high-resistance continuity tester.

The CT-1 comes factory assembled with a list of applications, instructions, a test lead and power cord for $16.95. The unit is hand held and designed for safety.

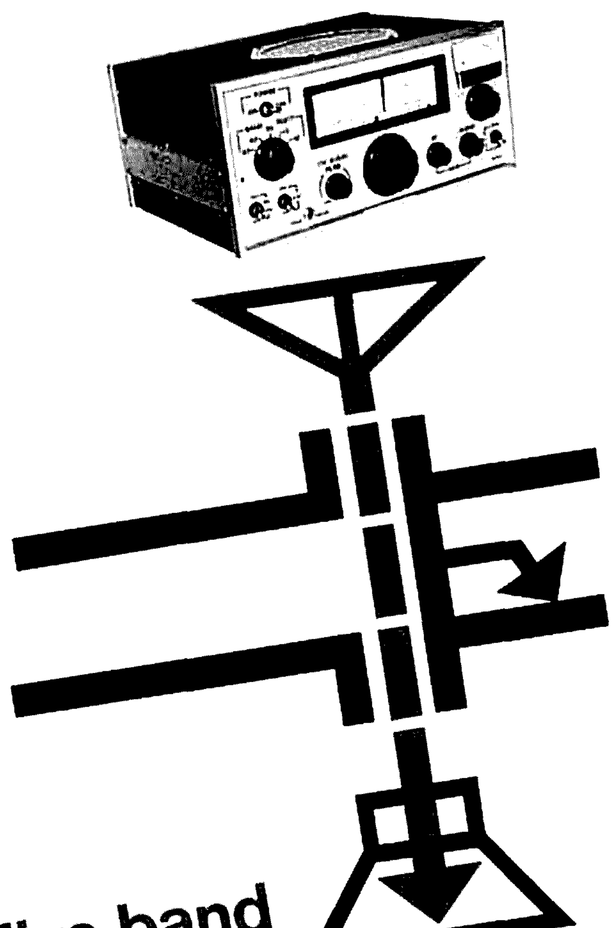
For a complete fact sheet write to Lee Electronics Laboratories, 88 Evans Street, Watertown, Massachusetts 02172 or use *check-off* on page 126.

75 cents

focus on communications technology . . .

ham radio magazine

JUNE 1972



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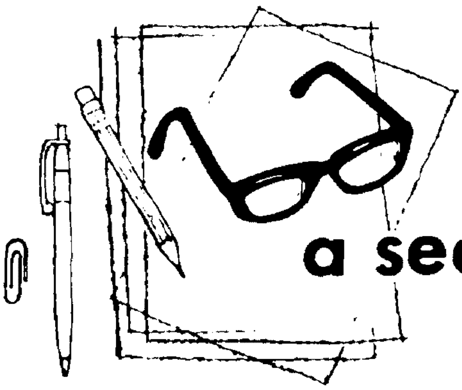


a second look

by jim fisk

Recently, I overheard a contact on 40 meters between two long-time amateurs who felt that amateur radio today just isn't as exciting, inviting and mysterious to today's youth as it had been to them forty years ago. They felt that in today's era of technological miracles, our hobby has less uniqueness and fewer opportunities to offer than it had in years past. Similar feelings have underlined many amateur discussions recently.

A. Prose Walker, W4BW, in speaking at amateur gatherings around the nation, has mentioned that amateur radio up to this day has been exciting and has seen tremendous technical progress and growth. But Mr. Walker maintains that the past has only been a prologue to what is to come.

Along with many others, we at *ham radio* note that the day of the amateur-built receiver may have passed in favor of the vastly superior and less expensive commercial version. However, look to the applications to which this commercial gear can be used for experimentation and growth.

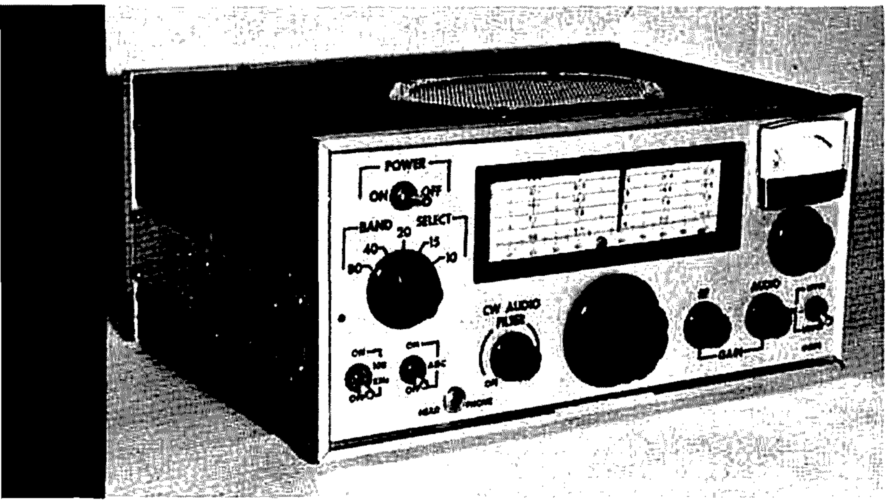
Some oldtimers say that even the thrill of working DX is gone now – anybody with money and a big antenna can work all the DX he wants. Yet today we see the phenomenal growth of QRP as hundreds of old-timers and experienced kilowatt-wielders leave their high-power equipment to marshal three of four watts and chase DX across oceans or across state lines.

Modern solid-state technology and manufacturing techniques have provided us with equipment which has fostered the amateur spirit – perfecting the art of getting the message through in spite of the limitations of equipment or power. Rather than making more "appliance operators," new low-power commercial equipment has presented new challenges and opportunities for fun and better training to help amateurs serve in the public interest.

A recurring theme of Mr. Walker's has been the undreamed of opportunities presented to today's amateurs by satellite communication. Rather than bemoaning the passing of the simple homebrew a-m transmitter, look toward the possibilities of intercontinental communication *when you want it* rather than at the whim of the ionosphere. The amateur's traditional communication expertise, inquisitiveness, patience, resourcefulness and determination must again come to the fore in this present and indescribably exciting field of amateur satellite communication.

Most importantly, though, our present level of sophisticated communications equipment gives us all the ability, and the real *responsibility,* to truly communicate with our fellow amateurs – either through the local fm repeater or on the other side of the world. And if that still isn't exciting, if it isn't challenging, if it isn't rewarding, if it isn't as new and vital as *today,* then maybe we should start the funeral dirge for amateur radio. However, I agree with Mr. Walker – the past is only a prologue to the future.

Jim Fisk, W1DTY
editor



five-band solid-state communications receiver

The A80-10 receiver — a straightforward reproducible design for 80- through 10-meter CW, ssb and a-m reception

M. A. Chapman, K6SDX, 1262 Orkney Lane, Cardiff, California 92007

This receiver, which I call the A80-10, has been designed for simplicity of construction, and uses standard solid-state circuits in a dual-conversion system. You can see from the photographs that an attempt was made to minimize the number of operating controls without compromising performance. So often, the extra knob (or switch) for some special feature is difficult to apply correctly, and many of them cause operator confusion.

The A80-10 receiver provides excellent amateur-band performance in any of the standard operating modes. The across-band sensitivity is less than 1 μV for 10 dB S+N/N ratio. The passband is controlled by a Collins 2.1-kHz mechanical filter, with added skirt improvement provided by an i-f transformer serving double duty as an impedance-matching network.

Frequency drift is negligible. Either of two methods is available to the builder to minimize drift problems — NPO capacitor compensation or the installation of a British *Tempatrimmer*,* an adjustable

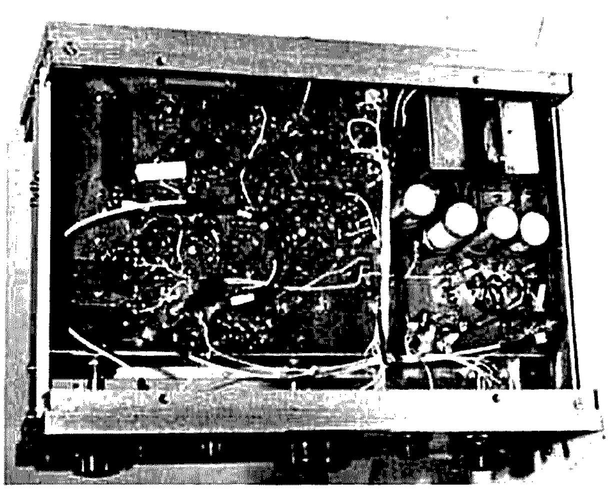
*The Oxley 6.5-pF *Tempatrimmer* is available from British Radio Electronics Limited, 1742 Wisconsin Avenue, N. W., Washington, D. C. 20007. The unit is priced at $13.25; add $.50 for postage and handling.

temperature-compensating device. Both methods satisfy today's stringent stability requirements.

The CW passband is effectively much less than that provided by the mechanical filter. An added op-amp audio filter allows peaking of any audio tone, thereby filtering out unwanted adjacent signals. A few minutes spent learning to operate the upper and lower bfo control with the CW audio filter should allow you to obtain passband selectivity that is adequate in the most dense operating conditions.

The A80-10 receiver has two poor functional features which should be pointed out. First, there are internal birdies at 7.0 and 21.0 MHz. The 7.0-MHz birdie is quite strong and can result in the loss of the bottom 2-kHz of the 40-meter band. Both birdies are second-overtone harmonics from the band-switching oscillator in the converter. The 21.0-MHz spurious is relatively small by comparison, but can be easily recognized. Second, the printed-circuit board, panel, etc., were not designed for a particular commercial cabinet.* All peculiar parts, panels, etc., were developed for this design: *on the plus side*, the size of the PC board and grouping of operating controls lends itself to a variety of cabinets at the builder's option.

Receiver performance will be directly proportional to the builder's patience and efforts. Poor wiring and assembly practice will result in disastrous failure. As in any design, compromise and options exist.



Rear view of the pc board.

table 1. Coil-winding data for the A80-10 receiver.

L1, L2	4 μH. 25 turns no. 32 on Miller 4500-2 coil form. Link is 4½ turns no. 32 on cold end. Resonates to 14.2 MHz
L3	40 turns no. 32 on Miller 4500-3 coil form. Link is 10 turns no. 32 on cold end. Resonates to 3.9 MHz
L4	1.2 μH. 10 turns no. 32 on Miller 4500-3 coil form
L5, L6	3.4 μH. 20 turns no. 32 on Miller 4500-3 coil form
L7	1.2 μH. 20 turns no. 32 on Miller 4500-3 coil form
L8	11 turns no. 32 on Miller 4500-2 coil form. Resonates to 24.5 MHz
L9	15 turns no. 32 on Miller 4500-2 coil form. Resonates to 17.5 MHz
L10, L11, L12	11 μH. 40 turns no. 32 on Miller 4500-2 coil form. Link is 10 turns no. 32 on cold end
L13	15 μH. 50 turns no. 32 on Miller 4500-2 coil form

Miller 4500 coil form is a slug-tuned ceramic form, 0.260" diameter. The −2 powdered iron slug (color coded red) is designed for the frequency range from 1 to 20 MHz. The −3 slug (color coded green) is designed for use between 20 and 50 MHz. Miller coil forms are available from the J. W. Miller Company, 5917 South Main Street, Los Angeles 90003.

The builder with a well-stocked parts reservoir will find a major portion of his receiver completed. The guy starting from scratch is up against it. Although there are a nearly infinite number of parts substitutions that can be made, only the gross types will be discussed.

basic design

An overall schematic/block diagram is shown in fig. 1†. Briefly, the receiver uses the 3.5- to 4.0-MHz band as the first conversion frequency, mixing with a high-frequency oscillator for a second conversion to 455 kHz. A 2.1-kHz Collins

*A complete set of drawings for the cabinet illustrated in the photographs is available from the author. Please include a large self-addressed, stamped, envelope.

†The power supply circuits, mixer circuit and 100-kHz marker oscillator are essentially the same as those published in the ARRL "Radio Amateur's Handbook," 1970 edition.

filter shapes the passband. The 455-kHz i-f signal is then amplified and beat against an upper or lower sideband oscillator in a product detector. Detected audio is amplified by a single IC to the 1-watt level, suitable for low-power speakers. Higher frequency bands are on a common PC board. Fig. 10 illustrates the functional arrangement for each section of the receiver. You can use your own judgement as to where to start. I found myself moving back and forth from section to section as components became available.

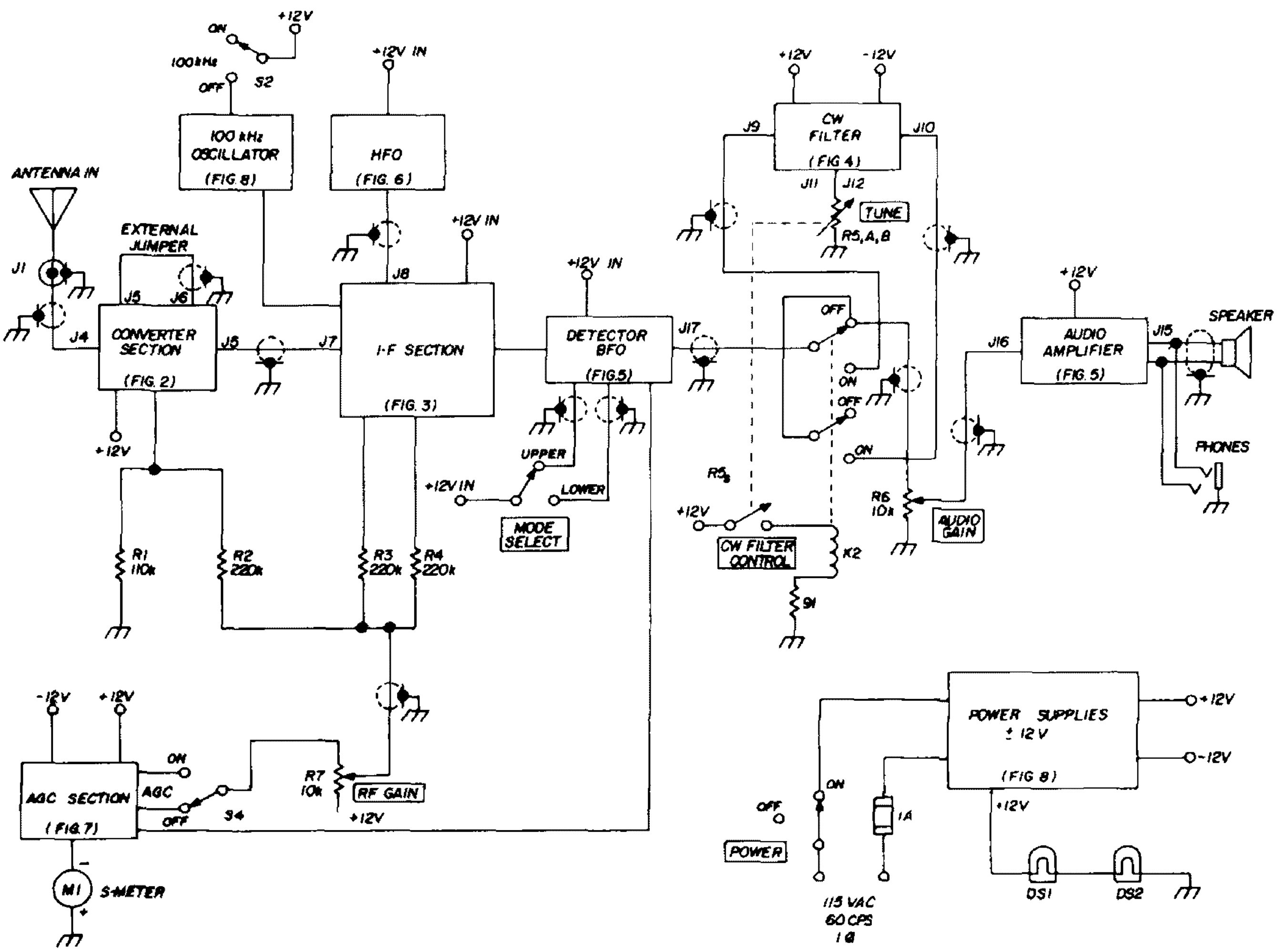


fig. 1. Block diagram of the all-band A80-10 high-performance communications receiver.

tuned with the simple two-stage crystal-converter method. Specific bands are selected from the front panel.

For CW operation, as previously mentioned, a tunable audio filter controls the detected signal. The same knob which is used for tone selection has a built-in switch which activates a relay circuit for in-out control.

construction

As you can see from the photographs, all of the active components are *mounted* on a common PC board.

The serious builder will be able to make an inventory of the items he needs and immediately recognize those items having long procurement times. I would suggest ordering the i-f transformers, filter capacitors and rotary switches immediately, even before the printed-circuit board is available.

The bfo crystal frequencies are determined by the 20-dB points of the Collins mechanical filter. It is important to obtain the filter before ordering the crystals. The bfo *and 100-kHz marker*

crystals are soldered directly to the PC board. Either ceramic or piston-type trimming capacitors may be used in the bfo and 100-kHz circuits.

If you like to do your own thing, the PC board layout in **fig. 10** has enough information for taping and etching your

construction sequence

The following sequence is a suggested procedure for building the A80-10 receiver:

1. Install all of the printed-circuit components, rf chokes and smaller filter components to the board.

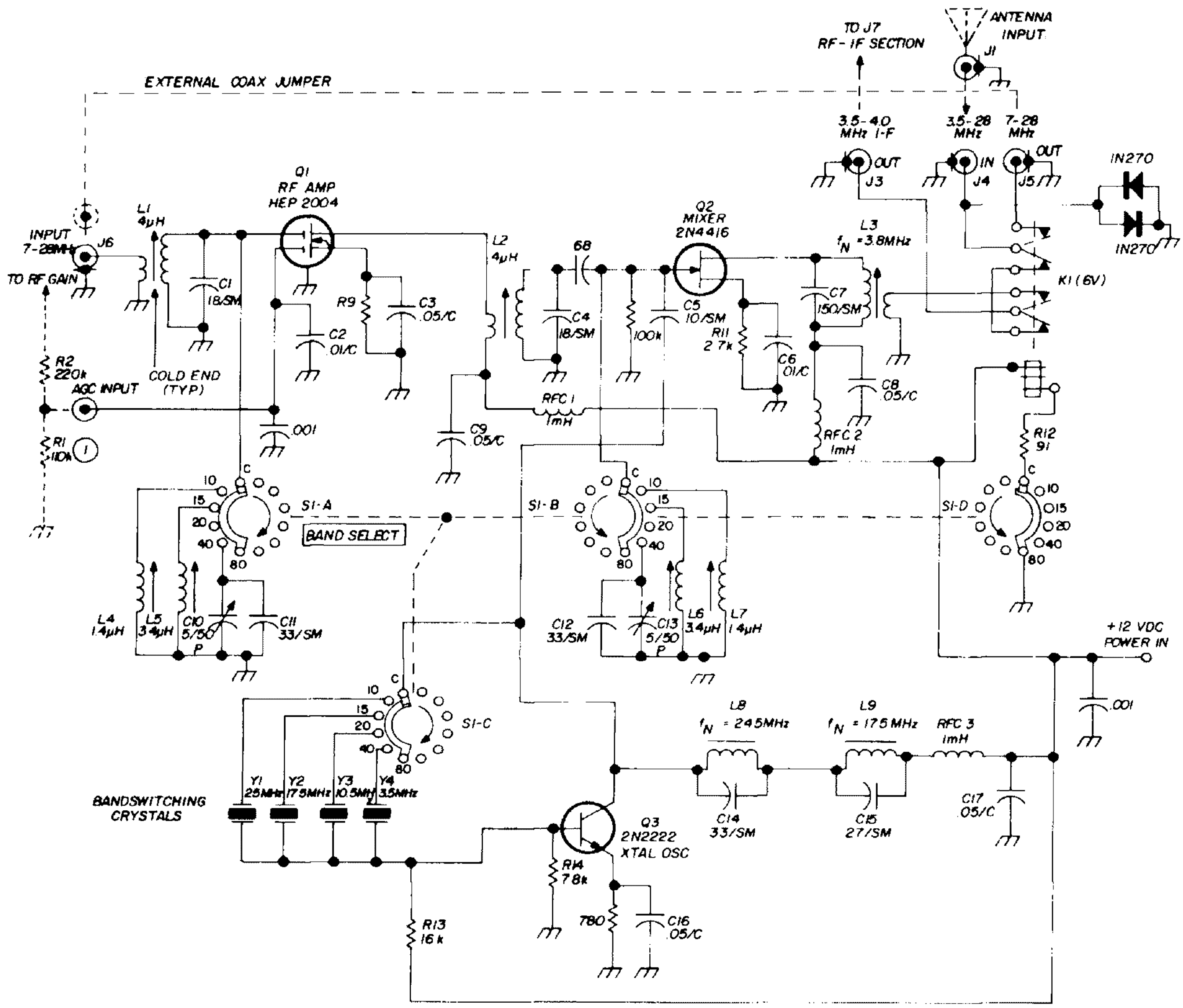


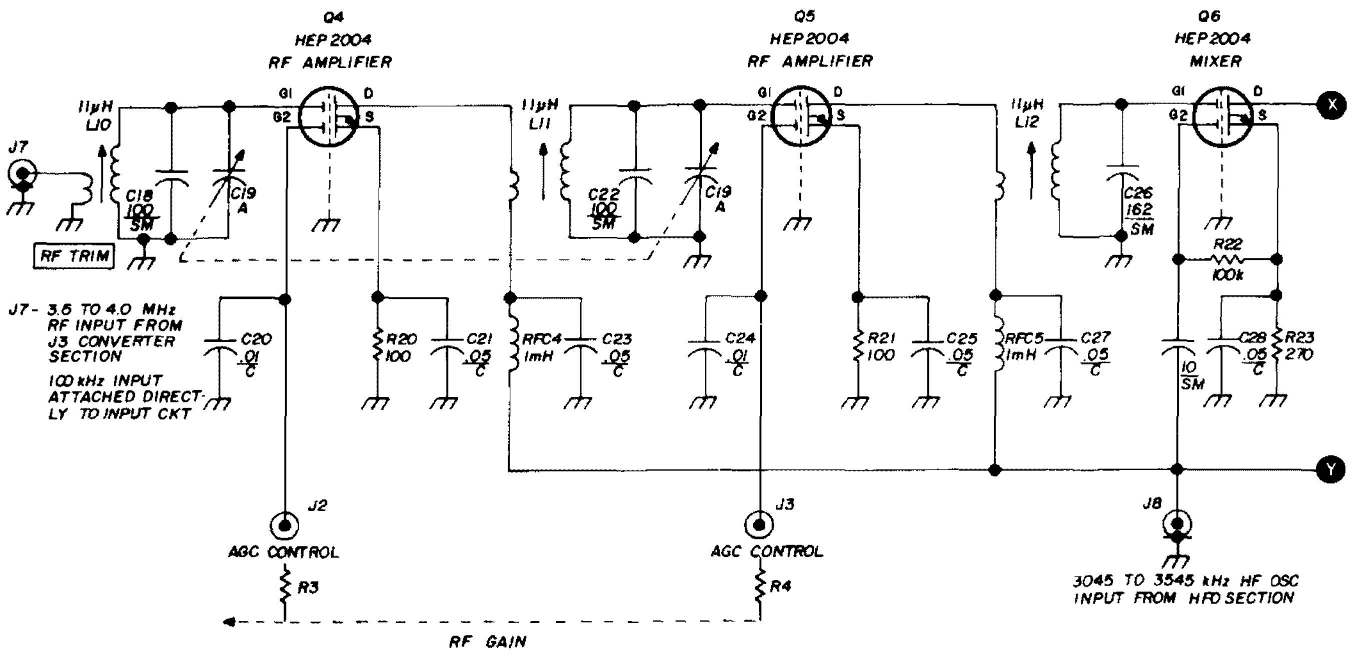
fig. 2. High-frequency converter covers the 40, 20, 15 and 10 meter bands. Complete coil-winding data is given in table 1.

own board, with whatever mechanical arrangements and component substitutions seem appropriate. However, **fig. 13** is a half-scale layout of the PC board used in my receiver. A double-sided board is used, with the same circuit pattern on both sides for stiff ground plane and continuity.

2. Add the dc power straps and agc lines.

3. Install the ICs, transistors and fets (with the exception of the dual-gate mosfets).

4. Add the inductors; each of these coils has been wound several times so the turns ratios are accurate and considerable lati-



tude exists. **Figs. 10** and **15** show the proper orientation of the inductor terminals.

5. Install the trimming capacitors.

6. Add the shields to the rf stage; 0.005 to 0.010-inch copper is suggested. If significantly thicker material is used, soldering to the PC board may be a problem (see **fig. 14**).

7. Add the shield to the converter stage.

8. Install the subpanel, rf peaking capacitor, bandswitch and the two relays.

9. Complete that portion of the wiring which interfaces with the subpanel components and the board.

10. Assemble the power-supply regulator panels and install them on the PC board along with the power transformers, rectifiers and filter capacitors. Add jumper wires to the regulator panels, but do not hookup the dc buses at this time.

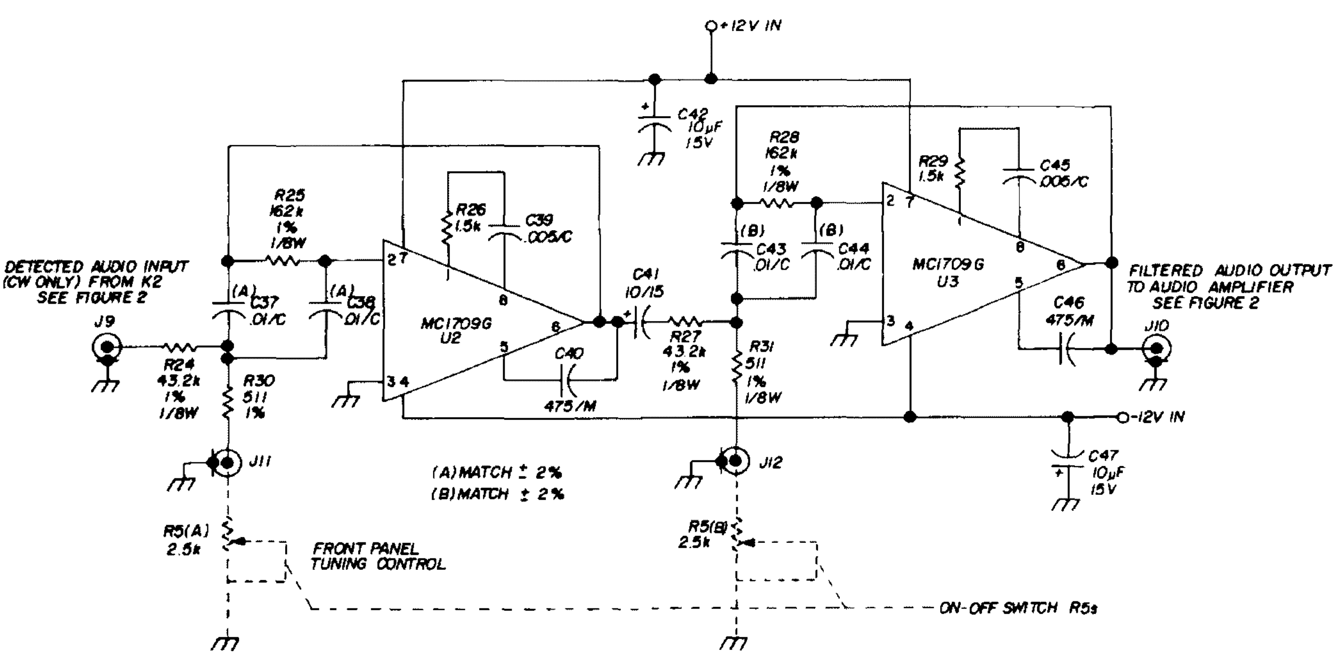


fig. 4. Audio filter provides excellent selectivity for CW communications.

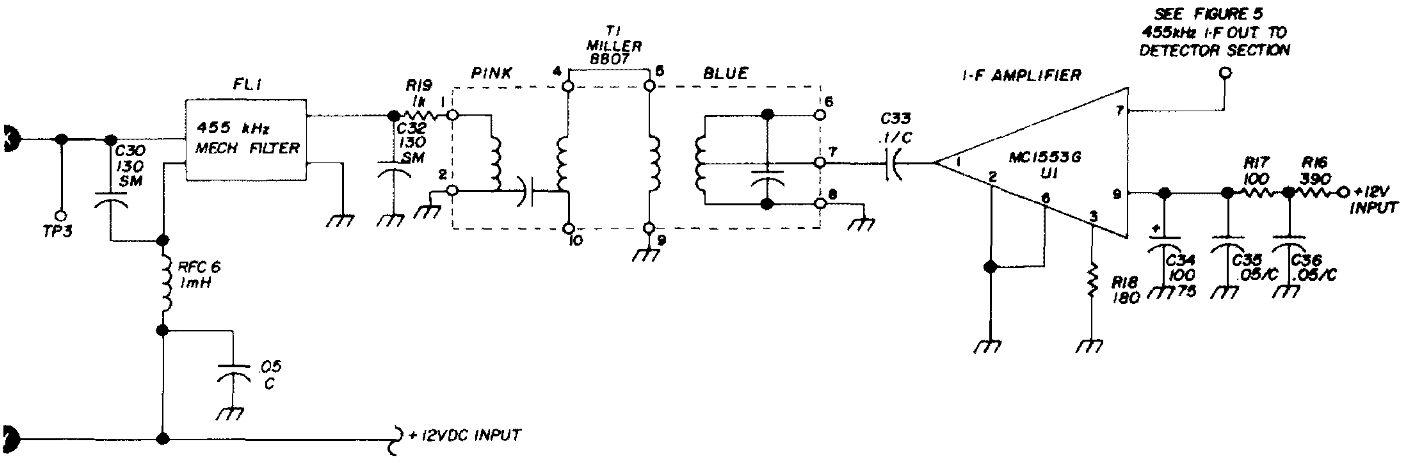


fig. 3. Circuit of the 3.5- to 4.0-MHz rf amplifier and 455-kHz i-f. The Collins mechanical filter provides desired selectivity.

11. Install the assembled printed-circuit board and subpanel into the cabinet. **Fig. 10** shows the brackets and supports which adapt the sub-assembly to the cabinet.

12. Add the front panel components and interwiring except for the S-meter wiring.

construction hints

The rf tuning capacitors and band-switch are all mounted on the same center-line. Use the switch mounting feet as a common point for mounting the PC board; the bushings are mounted on the subpanel as indicated in the cross-section

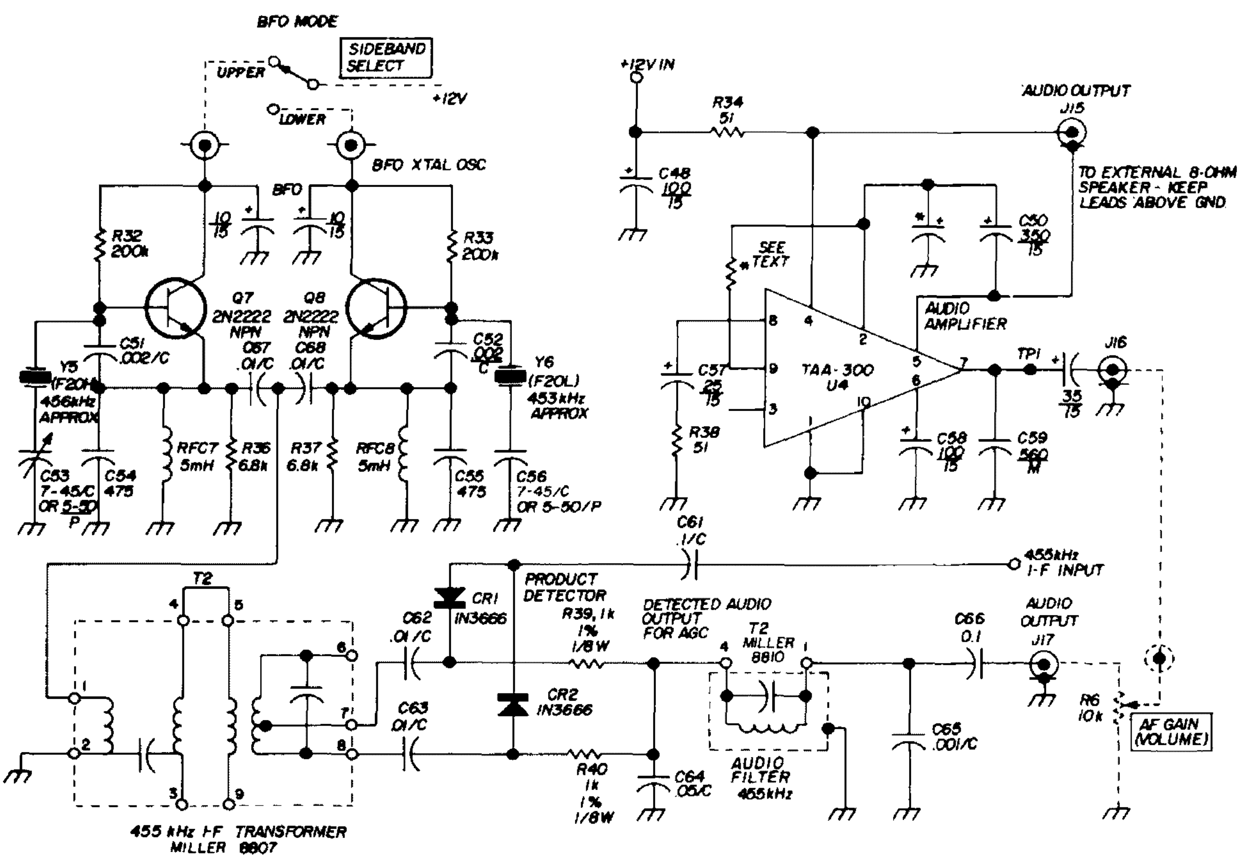


fig. 5. Bfo, product detector and audio amplifier circuits.

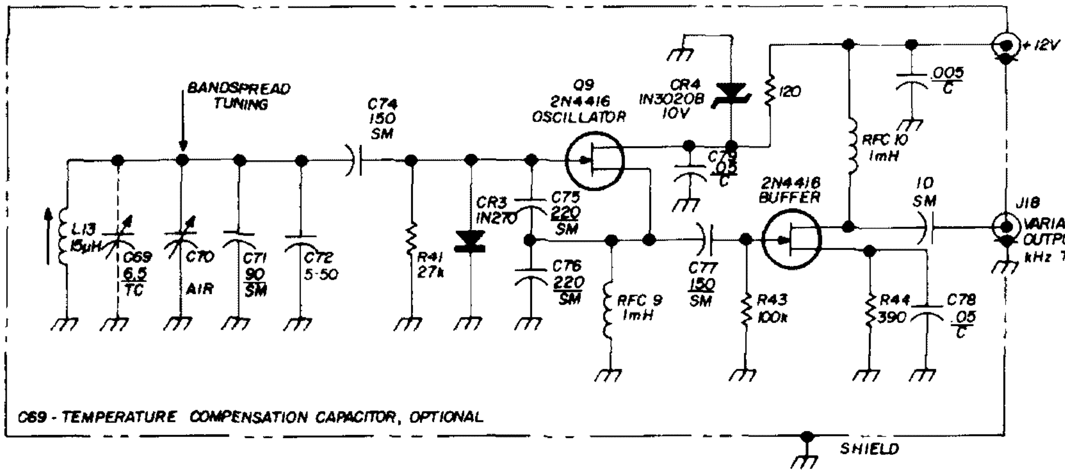


fig. 6. High-frequency oscillator circuit has excellent stability and low drift characteristics.

views in **figs. 11** and **12.** I found that the *feel* of the Centralab PA-2021 rotary switch was poor; there are ball-detent switches similar to the PA-2021 which you may want to use. Both Oak Manufacturing and Centralab have good miniature ball-detent units.

To make maximum room available for adding the jumper wires between the converter components and the bandswitch, remove all the excess contacts. This will minimize coupling problems as well. Referring to **fig. 10**, the shields between stages in the i-f and converter sections are just high enough to clear the ends of the inductors; a small bead of solder is run along the bottom of the shield as shown in **fig. 14.**

To provide for maximum slug adjustment in the inductors, the coils should be wound as shown in **fig. 15.** Complete winding data is given in **table 1.**

In my receiver I used no. 20 solid,

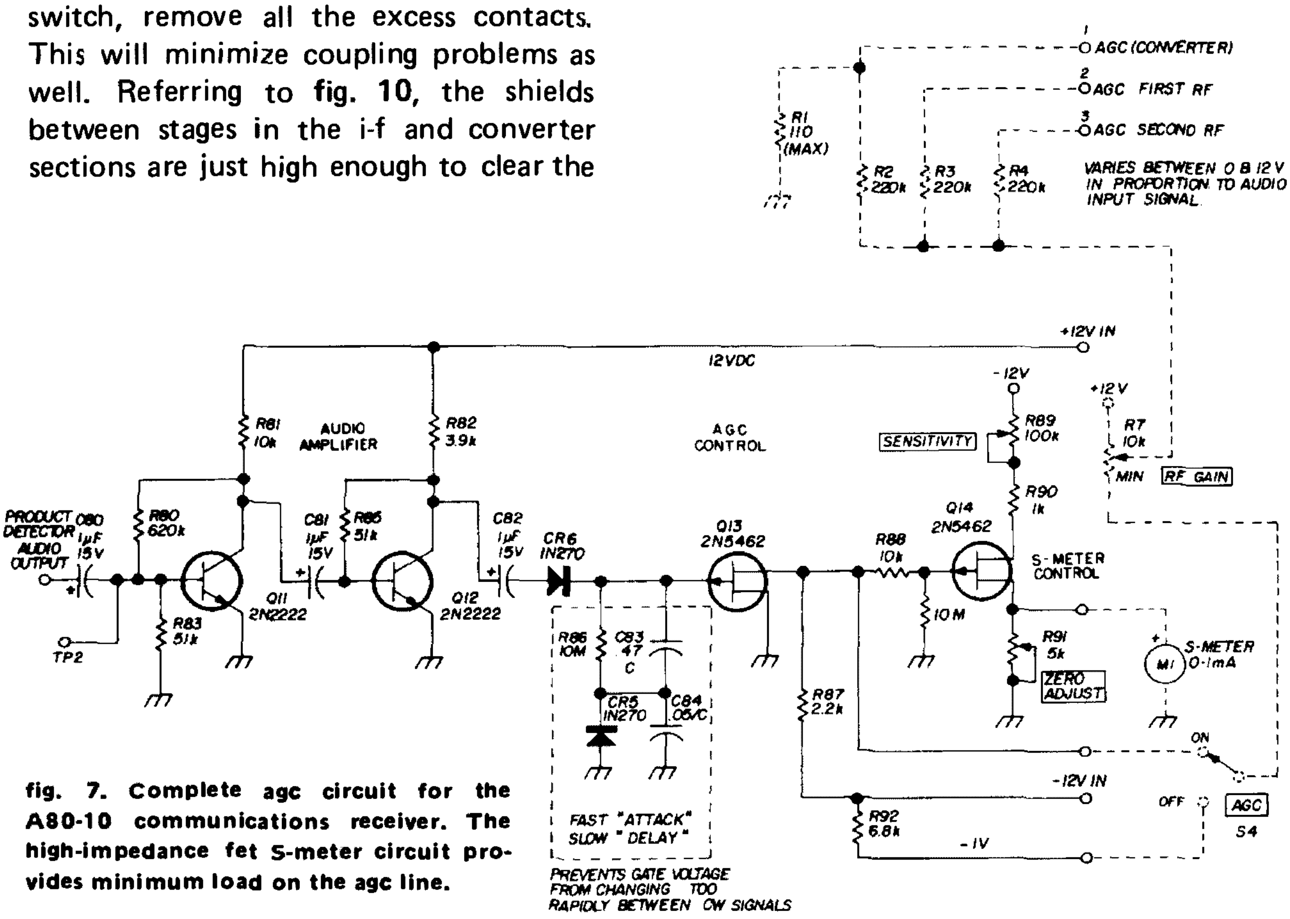


fig. 7. Complete agc circuit for the A80-10 communications receiver. The high-impedance fet S-meter circuit provides minimum load on the agc line.

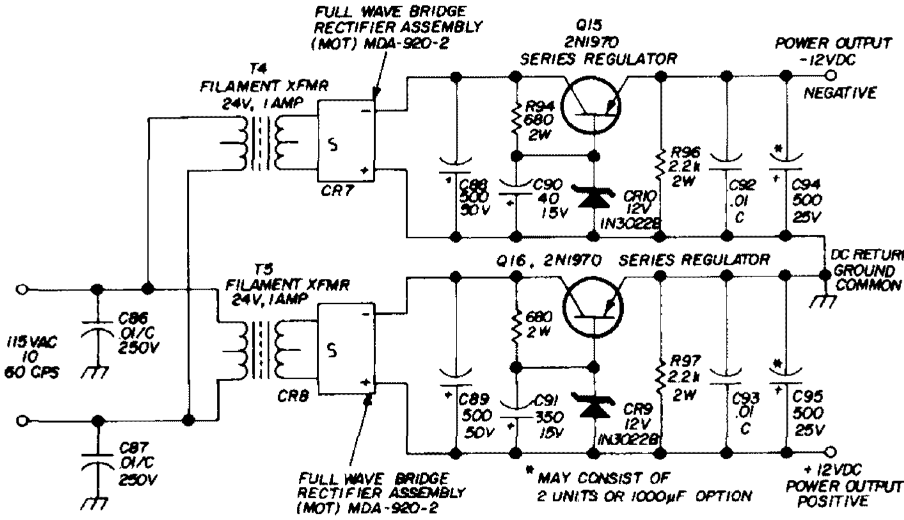


fig. 8. Regulated power supply for the receiver provides plus and minus 12 Vdc.

tinned wire for the jumpers from the tuning inductors to the printed-circuit board, bandswitch and resonating capacitors. For all other wiring I used no. 24; miniature coaxial cable was used for shielded leads.

If you look at **fig. 10** you will see that there is an offset between the integrated-circuit terminal centerline and the printed-circuit pads. The ICs should be installed as shown in **fig. 16**. The IC lead offset provides stress relief for the leads as well as allowing larger pad sizes to simplify etching the board. Note that the ICs are all installed from the back side of the board.

I tried HP2800 hot-carrier diodes in the detector, agc rectifier and agc attack circuits without significant improvement. These diodes provide low forward voltage drop and high leakage; however, the use of 1N270, 1N191, 1N914, 1N100 or 1N3600 diodes in these circuits provides almost equivalent performance. Don't be hesitant about using any good germanium device in the detector and agc attack circuits. Almost any good silicon rectifier will work well in the high-frequency oscillator circuit.

semiconductors

Transistor types used in the A80-10 receiver were standardized wherever possible. For example, all small-signal bipolar stages are handled by the 2N2222, an npn high-speed switching transistor. However, any general-purpose npn transistor may be used so long as it has comparable current gain (h_{FE} = 50 minimum) and adequate voltage ratings.

In the agc circuit, transistor Q11 is operated class A, but saturates with strong signals. Transistor Q12 operates almost entirely in class B to accentuate positive pulses and provide a wider gate voltage swing on Q13. Some adjustments in bias values may be necessary if substitute transistors are used in these stages.

The 2N5462 p-channel depletion-mode fet used in several stages is a

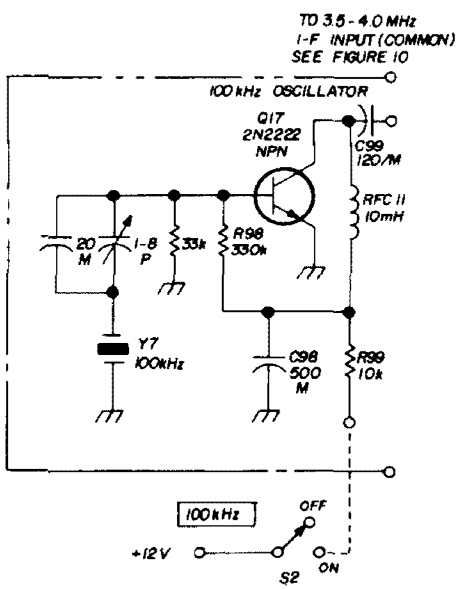
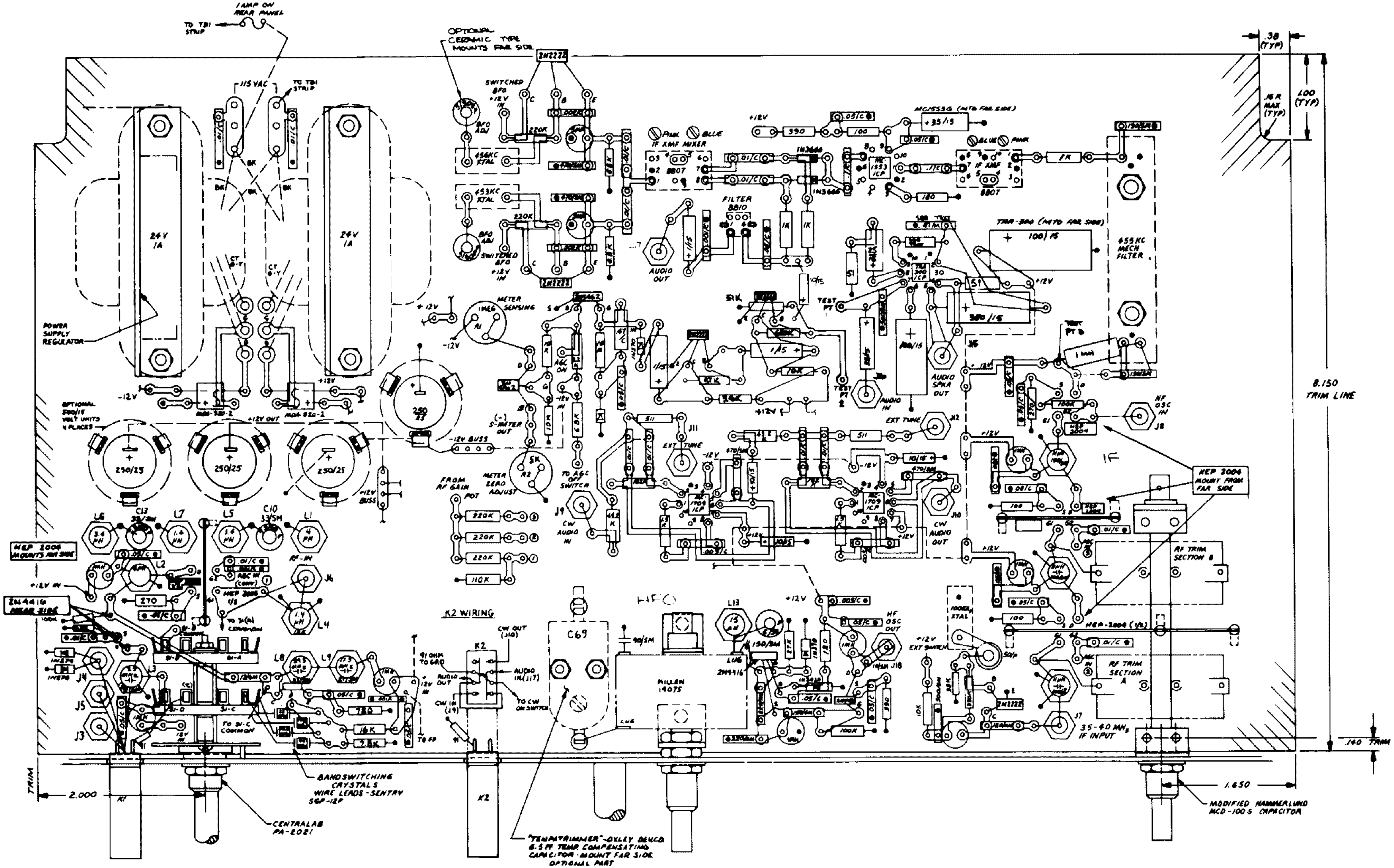


fig. 9. Circuit for the 100-kHz marker oscillator.


LAMP ON REAR PANEL
TO TB1 STRIP
115 VAC
OPTIONAL CERAMIC TYPE MOUNTS FAR SIDE
2N2222
SWITCHED BFO +12V IN
BFO ADJ
456KC XTAL
453KC XTAL
IF XMF MIXER
FILTER 8810
MC1553G (MTD FAR SIDE)
TRR-300 (MTD FAR SIDE)
455KC MECH FILTER
24V 1A
POWER SUPPLY REGULATOR
OPTIONAL 500/15 VOLT UNITS 4 PLACES
250/25
METER SENSING
S-METER OUT
METER ZERO ADJUST
FROM RF GAIN POT
TO AGC OFF SWITCH
CW AUDIO IN
CW AUDIO OUT
AUDIO OUT
AUDIO SPKR OUT
AUDIO IN
EXT TUNE
TEST PT
HF OSC IN
HEP 2004 MOUNT FROM FAR SIDE
RF TRIM SECTION B
RF TRIM SECTION A
HF OSC OUT
MILLEN 14075
100KC XTAL
3.5-4.0 MHz IF INPUT
HEP 2006 MOUNTS FAR SIDE
2N4416 REAR SIDE
K2 WIRING
C69
BANDSWITCHING CRYSTALS WIRE LEADS - SENTRY SGP-12P
CENTRALAB PA-2021
"TEMPATRIMMER"-OXLEY DEUCO 6.5 PF TEMP. COMPENSATING CAPACITOR - MOUNT FAR SIDE OPTIONAL PART
MODIFIED HAMMERLUND MCD-100S CAPACITOR
8.150 TRIM LINE
.38 (TYP)
1.00 (TYP)
.16 R MAX (TYP)
.140 TRIM
1.650
2.000
TRIM

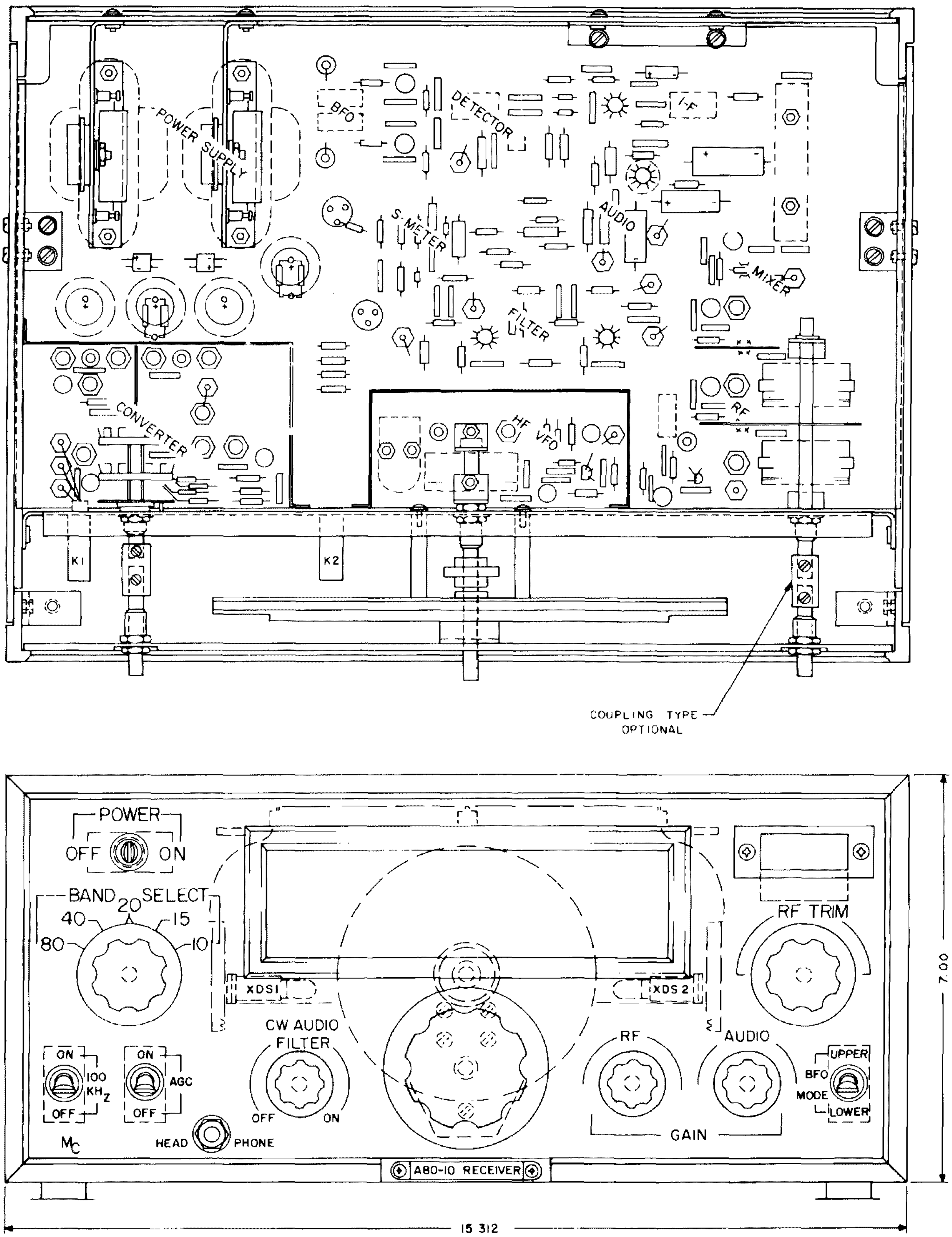


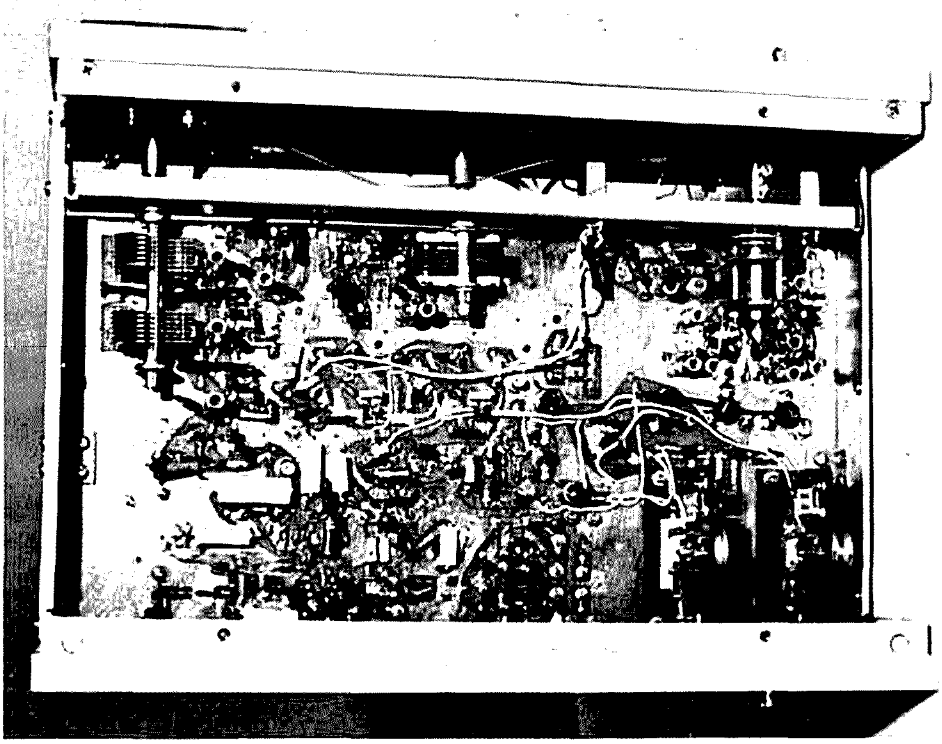
fig. 10. The complete parts layout for the A80-10 is shown on the facing page, the printed-circuit layout is shown in fig. 13 and the construction and panel layout of the receiver is shown above.

general-purpose type. The 2N5463 or 2N4360 are acceptable substitutes. Since large variations in gate voltage vs drain current between different devices will cause the maximum agc voltage swing to vary considerably, it's a good idea to have several transistors to select from. Adjust the 2.2k agc drain resistor slightly (between 1.5k and 3k) to maximize the agc voltage swing.

The 2N4416 n-channel fet used in the rf amplifier and mixer stages provides exceptionally good noise figure. Alternate types, which work well in these circuits, are the 2N3823, 2N5459, MPF-102, MPF-105 and MPF-108. However, the 2N3823 is the best choice for low mixer noise and maximum vfo output voltage.

The HEP-2004 dual-gate mosfet is used in the vhf amplifier and mixer circuits. The MFE-3006, MFE-3007, MFE-3008, MPF-120, MPF-121 and MPF-122 are superior substitutes since wider gain variations can be expected from the HEP-2004. However, there is sufficient i-f gain to offset this, and the cost of the HEP-2004 is much less than the recommended substitutes. Protective diodes are installed across the antenna input circuits for over-voltage protection.[1,2]

A Motorola MC-1553G wideband linear amplifier IC was selected for use in the A80-10. The MC-1552G may be used with somewhat reduced gain and sensitivity. I experimented with some similar ICs manufactured by RCA, but they all required an additional buffer stage to obtain equivalent results. The cost of the MC-1553G is relatively high, but well worth the investment because of the circuit simplicity it provides.



Top view of pc board with converter and hfo covers removed. The i-f is along the left, the hfo is in the middle.

Although I used the Motorola MC1709C operational amplifier in the audio filter, a variety of suppliers makes this same device, including Fairchild (W5B77092IX), Philco (PA70931), Raytheon (RM4709) and Texas Instruments (SN52709L). In a recent issue of *ham radio,* WA1JSM discussed some additional design features of this device, as well as applications.[3]

The Amperex TAA-300 audio amplifier IC used in the audio power amplifier stage provides 1-watt into an 8-ohm load. This device requires a heat sink. You may

want to obtain a copy of the Amperex data sheet on this device since there are considerable variations in external capacitors and feedback resistors for various operating *requirements.*[4]

In some installations, to prevent instability, you may have to connect a capacitor (47 pF minimum) from pin 2 of the TAA-300 to ground. The printed circuit board indicates this component location. However, motorboating may result if this capacitor is installed prior to test and found not to be required. Some devices may require an additional capacitor (100 to 500 μF) from pin 4 to ground to prevent low-frequency oscillation.

test and alignment

Before applying any power, double check all your wiring connections, checking for continuity and short circuits. Check all transistors, capacitors and ICs for correct orientation. Without the power-supply regulator section connected to the circuit, you should obtain the following resistance readings from the plus and minus 12-volt lines to ground:

+12V bus approximately 50 ohms
−12V bus approximately 400 ohms

These measurements are made with the dual-gate-mosfet transistors installed in the board.

Apply ac power to the unloaded power supplies. The output voltage should be between 11.2 and 12.2 volts. Variations in zener-diode tolerances cause wide variation in output voltages. If the power supply voltages are within tolerance, connect the ±12-volt outputs from the regulator circuit to the corresponding bus points on the printed-circuit board.

Apply a 5- to 15-mV, 1000-Hz audio signal to TP-1. If the speaker is disconnected, install a 10-ohm resistor across the two output leads; connect a scope across the resistor — the audio signal will be displayed on the scope if everything is working properly. Of course, if the speaker is connected, you will hear the 1000-Hz tone.

Disconnect the speaker and install a 10-ohm resistor in its place. To properly test the agc circuit all of the rf stages must be tested and aligned. For preliminary agc alignment, however, apply a 25- to 50-mV audio signal to the base of Q11; varying this input voltage should cause a corresponding variation in negative voltage at the drain of Q13. If all is going well, a zero audio level applied to the base of Q11 should result in a corresponding Q13 drain voltage of −1 V; increasing the audio signal level to 25 or 50 mV should cause the Q13 drain voltage to change proportionately to −6 or −9 volts.

If the Q13 drain voltage swing is not as large as indicated, then something is wrong. Since Q11 is operating in class A, the collector voltage should be about 6 V, and the output signal should be nice and clean except when overdriven — then the output should start saturating a little. If Q12 is operating correctly the collector voltage should be about 1 or 2 volts with no signal, and 6 to 10 volts with a high input and some saturation may be apparent; i. e., the output could consist of square waves at about 6 to 10 volts positive.

The gate voltage on Q13 should be slightly less than the collector voltage of Q12, and vary in direct proportion. With this much gate voltage swing the drain current should cause the previously indicated voltage variations across the 2.2k resistor. I had to play a little with this circuit since there are so many variables between different fets that under some circumstances Q12 can operate in class A. Just a few volts on the gate of Q13 will provide a significantly large agc voltage. The circuit in fig. 7 gives a little larger gate voltage swing under touchy CW conditions where the large agc voltage will save your eardrums.

Another agc circuit which works well if the beta of the transistors is high is shown in fig. 17. Note that the value of the coupling capacitors between stages is unimportant so long as you use a minimum value of 0.1 μF. If higher values are used as *suggested in fig. 7,* there is an improvement in low-frequency response. Diode CR6 can be any device exhibiting

low forward voltage drop and moderate to high leakage such as the 1N914 or 1N191. Other circuits you might want to try are shown in Bill Orr's "Radio Handbook."[5]

s-meter

The 1970 ARRL "Radio Amateur's Handbook" suggests a typical S-meter circuit for receivers which use negative agc lines. However, the input impedance of the circuit is so low it severely loads the agc line. The fet S-meter circuit I used in the A80-10 (see **fig. 7**) is similar to the ARRL circuit, but does not load the agc line.

Adjustment of the S-meter circuit is quite simple since the sensitivity can be set for any preferred rf reference level. The meter responds to the decay agc feature so it does not hang up on a-m signals.

high-frequency oscillator

If this circuit is properly wired, and the coil is properly wound, there is absolutely no way that it will not work (see **fig. 6**). To align the oscillator, first set the bandspread capacitor to the center position. Grid dip the coil to the same approximate frequency (3.25 MHz) and apply power. Signal output should be from 1 to 3 volts peak, depending on the transistor used in the circuit.

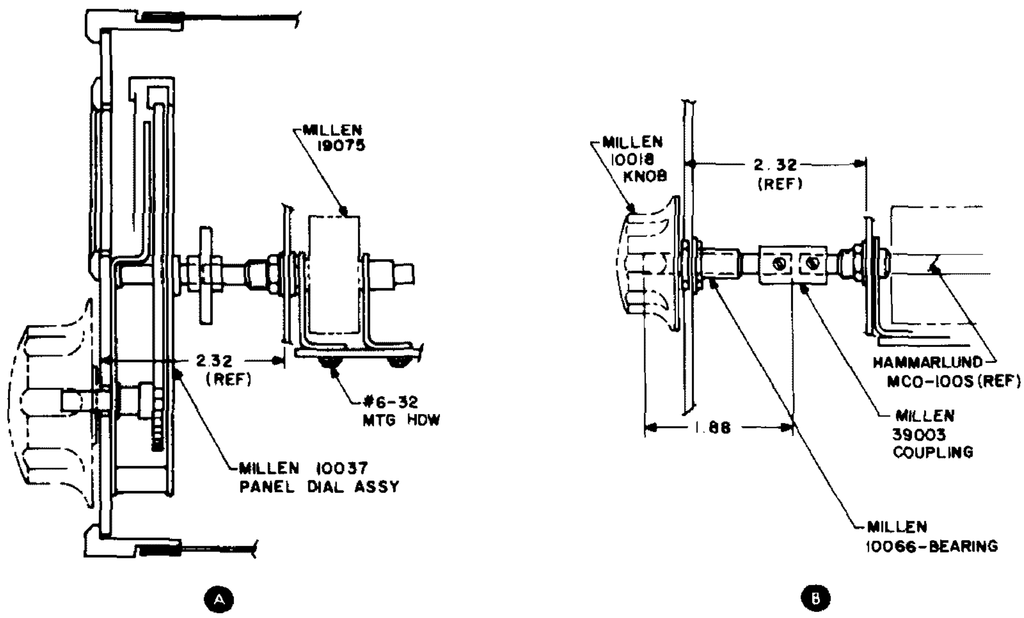


fig. 12. Construction details of the main tuning dial (A) and the rf trimming capacitor (B).

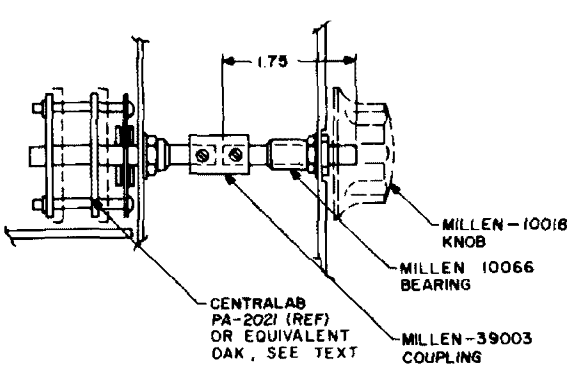


fig. 11. Construction details of the band-selection switch.

The oscillator may be built without the *tempatrimmer* if you wish. If the *tempatrimmer* is not used, the 100-pF silver-mica capacitor should be an NPO type to compensate for temperature drift. The voltage of the zener diode is unimportant – 8, 9 or 10 volts is satisfactory. If you use an 8- or 9-volt zener, slightly *increase the series resistance.*

The L/C ratios used in the oscillator will allow you to adjust the oscillator from 350 to 500 kHz on five bands. This

means that if you are particularly interested in a certain segment of the bands, by properly selecting the converter crystals and adjusting the L/C combination, bandspread of a particular band segment can be adjusted to suit your interest.

Let's assume that you are interested in a straightforward 500-kHz bandspread as results will provide temperature drift levels less than 25 Hz hour.

Probably the easiest i-f section to build and align is the one shown in **fig. 3**. The coils are easily wound with an error latitude of one or two turns. The coil/capacitor assembly is shown quite clearly in **fig. 10.** Put the trimmer capacitor (C19) in the half-mesh position, grid dip

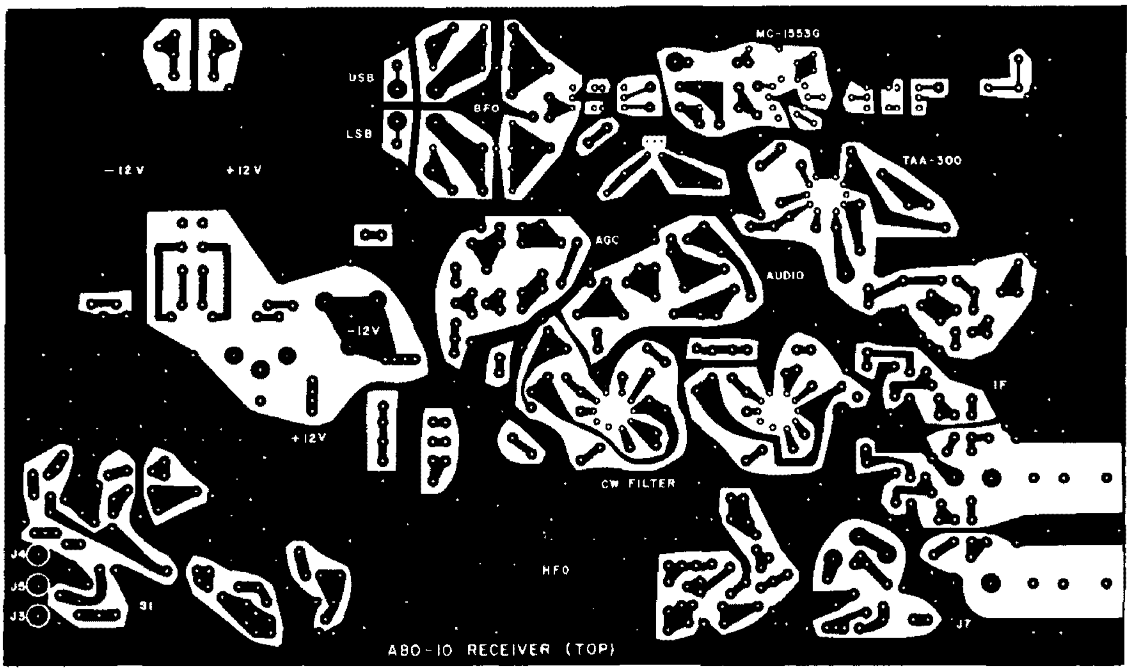


fig. 13. Printed-circuit board layout for the A80-10 communications receiver. Full-size templates are available for 25c from ham radio. Printed-circuit boards with critical component holes drilled are available from Spectrum Research Laboratory.*

illustrated. Not too much trial and error is required to obtain low end frequency output at 3045 kHz with maximum capacitance and minimum slug; then tune to the other end of the scale, setting the inductor slug to a frequency slightly lower, and then backing off the capacitor. Returning to the low end and trimming both inductance and capacitance should get the oscillator into the required tuning range (3045 to 3545 kHz) within 4 or 5 tries.

Let the receiver operate for 72 hours, retrim the tuning components, and then coil-dope the L13 slug and capacitors into position. Adjustment and installation of the *tempatrimmer* is best accomplished using the manufacturer's instructions; the the inductor slugs to 3750 kHz in each of the rf and mixer stages. Recheck circuit wiring and make sure that agc voltage is available at the mosfet gates.

Apply a 5-mV signal to the output of the 455-kHz mechanical filter or at TP3. If the i-f amplifier is operating correctly, an oscilloscope will display an audio signal at the audio input capacitor or across the 10-ohm load resistor. If you have finished the agc and S-meter sections, then visible S-meter movement

*Printed-circuit boards are available from Spectrum Research Laboratory, Inc., P. O. Box 5824, Tucson, Arizona 85703. The double-sided plated, completely drilled board is $27.50, including handling and postage.

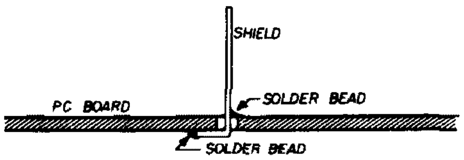


fig. 14. Installation of shields on the printed-circuit board.

should occur and agc control should be apparent by putting the agc switch in the *on* position and varying the 5-mV 455-kHz input signal from 1 to 15 mV while monitoring the mosfet gate voltages.

Install the dual-gate mosfet in the rf and mixer stages. Connect the hf oscillator to the mixer. Apply a 30-μV 3750-kHz signal to the input connector of the i-f and attach the oscilloscope to the output of the i-f amplifier IC. Tune the hf oscillator to a point where an i-f output is apparent. Adjust both slugs of the i-f transformer for maximum output signal. Adjust coils L12, L11 and L10, in that order, for maximum output signal. Now go back and readjust all trimmer components for 3750 kHz, and zero adjust the band-spread and capacitor trimmer. At this point, the receiver may be used for 80-meter reception.

To verify receiver operation, apply a 1-μV signal to the input with an oscilloscope monitor at the audio output. Tuning across the 1-μV signal and peaking the trimming controls should produce an obvious audio signal. With a 30-μV signal applied, 100% modulated with 1000 or 400 Hz, the upper and lower sideband trimming capacitors in the bfo circuit can be adjusted to provide audio reception at the appropriate level. Individually tune the bfo capacitors for maximum audio output. The output levels can be matched by adjusting R32 and R33.

cw filter

Properly wired and with good op-amps, the CW filter circuit should function immediately. Review the wiring of relay K2; turning the CW control clockwise should activate the relay, switching the audio through the operational amplifier circuits rather than directly to the audio amplifier. The potentiometer provides tunable audio; for CW operation a particular beat-note may be accentuated over another adjacent signal, much like a Q-multiplier. CW operators will find that by carefully using either the upper or lower bfo positions and the CW audio filter, passband characteristics will be equivalent to a 500-Hz filter. If you don't want to use 1% precision resistors, matched carbon composition types with values close to those indicated in fig. 4 are quite suitable for this circuit.

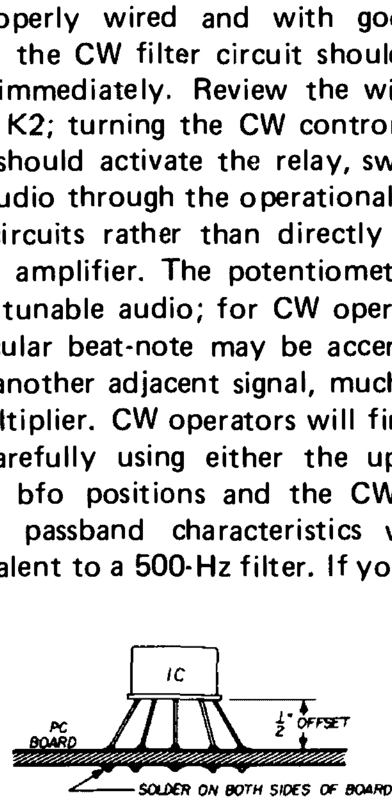


fig. 16. ICs are slightly offset as shown here to provide stress relief.

oscillators

Any npn transistor with current gain greater than 50 should work without difficulty in the bfo and 100-kHz oscillators. In the 100-kHz oscillator, substitution of a inductance value other than 10 mH will cause odd collector loading and erroneous oscillator frequencies. Adjustments in the collector-to-base resistor R_A will correct for variations in device gain and crystal impedance (see **fig. 18**).

converter section

To align the converter, the following procedure is suggested. First, verify the operation of transistor Q3, the crystal oscillator, moving the switch from the 40-

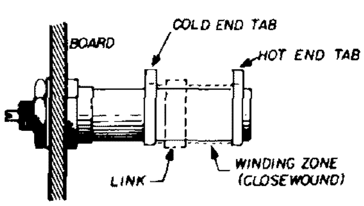


fig. 15. Inductor assembly. Complete coil winding data is given in table 1.

through 10-meter positions. Oscillator operation should be apparent by monitoring the common terminal of S1-C for the signal; the voltage level may vary between crystals. Some difficulty may be experienced in obtaining stable oscillation on the correct frequency in each of the oscillator positions. Variations in impedance between crystals *will* cause the base bias voltage to shift and cause erroneous output. Some adjustment of R13 and R14 is required, depending upon the device you use. In the circuit in **fig. 2**, the gain of the transistor was approximately 85. VE3GFN suggests a slightly different circuit;[5] however, I was not able to duplicate his circuit and obtain a stable output signal in all four crystal positions.

Minimum output voltage in the 10-meter position is approximately 1 volt.

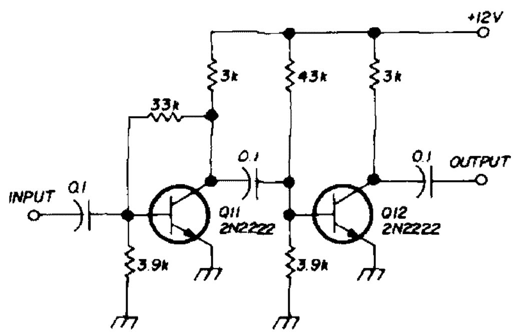


fig. 17. Alternate agc circuit.

Inductors L8 and L9 should be peaked for maximum signal output level at the *appropriate crystal frequency*. Grid dip L1 and L2 for 14.2 MHz, and L3 at 3.8 MHz. Put the band selection switch in the 20-meter position and apply a 100-μV 14.250-MHz signal to the antenna input. With a scope monitoring either the i-f amplifier output or the audio output, tune the hf oscillator for apparent signal reception. Adjust L3 for maximum signal output. Apply a 10-μV signal to the antenna input and adjust L1, L2 and L3 for maximum output level.

Put the band selection switch in the 40-meter position and apply a 100-μV 7.250-MHz signal to the antenna input. Adjust only C10 and C13 for maximum output signal.

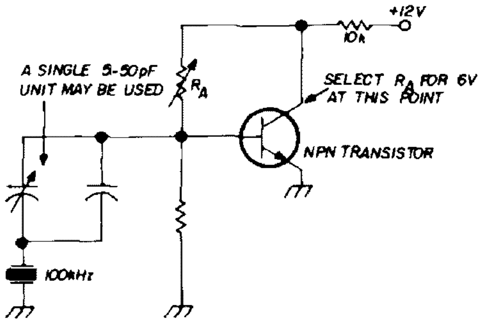


fig. 18. When initially adjusting the bfo or 100-kHz oscillators using other than the recommended transistors, it may be necessary to select the base-to-collector resistor R_A to compensate for variations in transistor gain and crystal impedance. Adjust R_A for 6 volts at the collector of the transistor.

The 10- and 15-meter bands are aligned in the same way, peaking the appropriate inductors for maximum signal. Final trimming of all inductors should be made again after installing shielding or covers around the converter section.

Some sensitivity and gain losses can be expected on 10 meters if you substitute different devices. Substitution of HEP 2004 transistors and adjustments in the value of R1 may be necessary. The more stable MFE-3006, MFE-3007 or MFE-3008 might prove useful.

And last, but not least, adjust transformer T2 for minimum 455-kHz speaker hiss.

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ham radio

integrated-circuit sequential switching for Touch-Tone repeater control

Modular construction combined with three-digit sequential decoding gives this all-IC unit flexibility with immunity from accidental keying

R. B. Shreve, W8GRG, 2842 Winthrop Road, Cleveland, Ohio 44120

Touch-Tone signals, which combine one frequency below 1 kHz with one above to represent each of ten digits, can be decoded with filters and relays or with integrated circuits. The Bell System 247B KTU uses LC filters, transistors and relays; a solid state IC decoder is described in the July, 1971 *QST*.[1] Either will provide a discrete digital impulse for each Touch-Tone signal received.

Using one of these impulses to control a switched circuit is easy — so ten tones should make it possible to perform ten functions. And they would, if no one else ever transmitted the same tones on the frequency you selected for a control frequency. To make a system that will act only on signals addressed to it and disregard all others, circuits are needed that will recognize multiple digits in a prescribed sequence. This article describes such a system.

design requirements

What should a Touch-Tone control unit for a repeater provide? At the Lake Erie Amateur Radio Club repeater, WB8CQR, the objectives were:

1. Accept Touch-Tone signals as reproduced at the output of a standard fm receiver.

2. Retransmit signals not addressed to the repeater control, but otherwise disregard them, so that others can use Touch-Tone on the same frequency

without interfering with the repeater controls.

3. Recognize signals directed at the control unit, interpret them and act on them without retransmitting them.

4. Disregard erroneous, irrelevant or incomplete codes, resetting to its stand-by state automatically.

The control unit is designed for a repeater with the transmitter and the primary receiver about a half mile apart, connected by a single telephone pair. There are two receivers on the primary input frequency, one at the main receiver site and one remote, both tone guarded. A second receiver at the main site, on a secondary frequency and unguarded, is used principally by base stations.

Wire control of the transmitter from a number of remote points is provided in accordance with the station license. The *Touch-Tone* control unit, located at the main receiver site, supplements the wire control but does not replace it. The Touch-Tone unit performs the following functions:

1. Disables the keying circuit to the transmitter, without affecting the receivers.

2. Switches the output to an alternate frequency.

3. Switches a guarded receiver from tone guarded to open.

4. Disables each receiver separately. The disabling circuits are interlocked so that it is impossible to disable all inputs at the same time.

Since each of the listed functions must be reversible, a total of twelve codes is required. Although two-digit codes would suffice, and allow plenty of room for expansion, it was felt that three-digit codes would give enough greater security against false or accidental keying to justify the additional equipment.

To get greatest flexibility during development and maximum adaptability to future change, all integrated circuits are grouped in modules, on etched circuit boards. The boards are approximately four inches square and plug into sockets

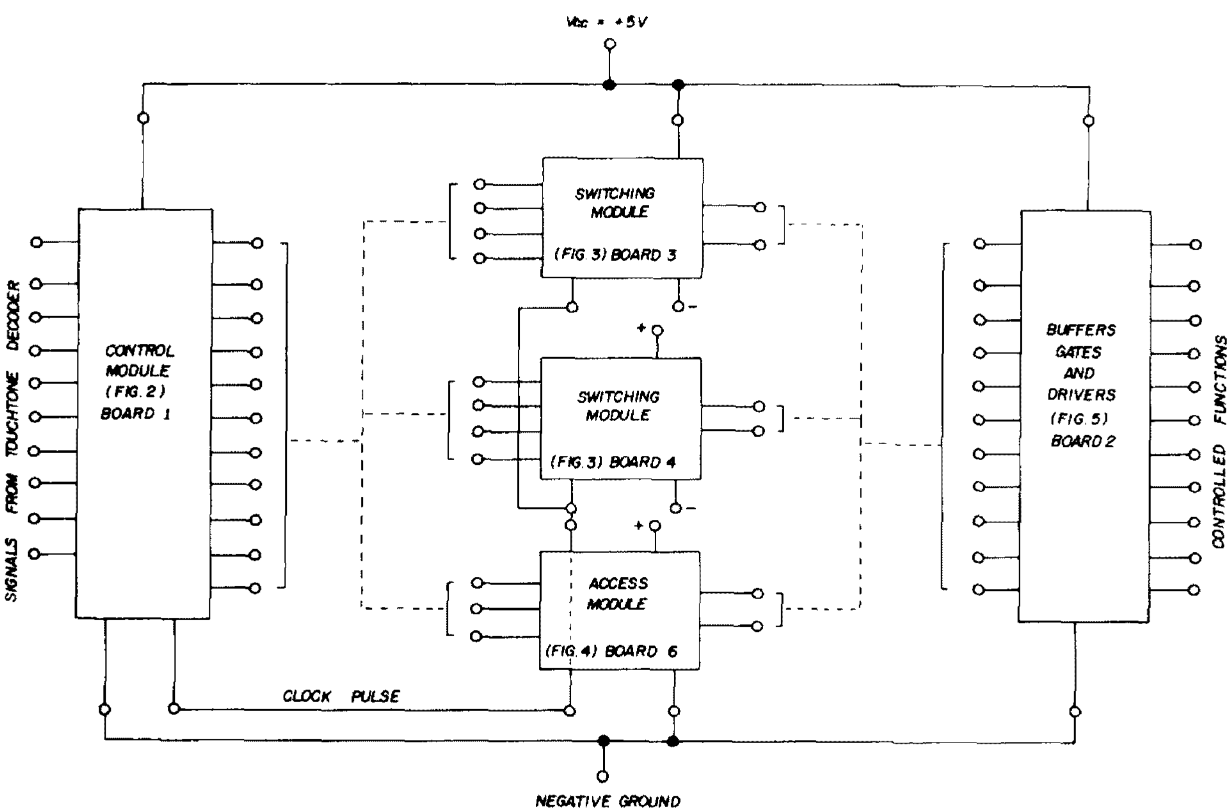
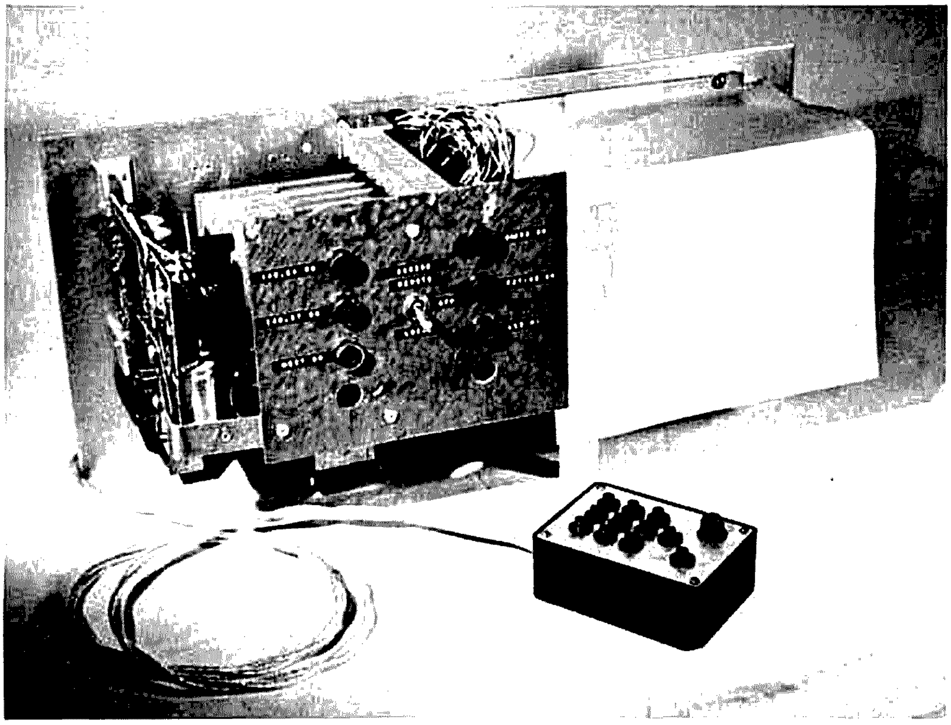


fig. 1. Overall block diagram showing relationship between all modules.

The unit occupies one half of a 19-inch rack. On the left is the power supply; the tops of the logic boards are visible to its right behind the panel on which indicator lights show the status of the six functions controlled. Space is provided for two additional lights if another switching module board is added later. The box in the foreground is the local control unit, with ten digital keys in the Touch-Tone pattern, a button to key the transmitter and a light showing when the gate is open to permit access to the unit.



that provide 22 possible connections to each side of the board. There are four types of modules, as shown in fig. 1. They are:

Control module (board 1, fig. 2). Processes incoming Touch-Tone signals and generates a clock pulse to drive the J-K flip-flops on the other circuit boards.

Buffers, gates and drivers (board 2, fig. 5). Provides additional logic elements to expand and combine outputs of other circuits.

Switching modules (boards 3, 4 and 5, fig. 3). Identical switching modules, two on each board; each module performs one on-and-off function.

Access module (board 6, fig. 4). A special switching module and timer that controls access to the switching modules.

Audio input to the decoder is taken from a circuit that also carries the transmitter audio input, so that tones heard by the decoder are also retransmitted if the transmitter is keyed. The decoder is one that grounds one of ten output terminals for the duration of each tone combination representing a digit. If the Blakeslee solid-state decoder is used, output inverters should be used, or the NOR gates in the output circuits should be replaced with POSITIVE OR gates, so that each Touch-Tone digit is represented by a "low" decoder output.[1]

control module

The control module circuitry is shown in fig. 2. Decoder outputs are connected to terminals 2, 4, 6, 9, 11, 13, 15, 17, 18 and 20. For each input an inverted output is provided at the adjoining terminal. These outputs are the digital impulses used to drive the sequential switching circuits. Eight of the digits are also connected to the inputs of an SN7409 quadruple AND gate used to expand the input capability of one half of an SN7413. The other two digits are connected directly to the SN7413 inputs — which digits to which inputs is immaterial; the arrangement shown was the simplest circuit-board layout.

The Texas Instruments SN7413 is a dual quad-input NAND gate with Schmitt-trigger action, characterized by hysteresis of about 0.8 volts between the levels of the turn-on and turn-off input signals. This makes it possible to operate one half of the gate as a multivibrator by

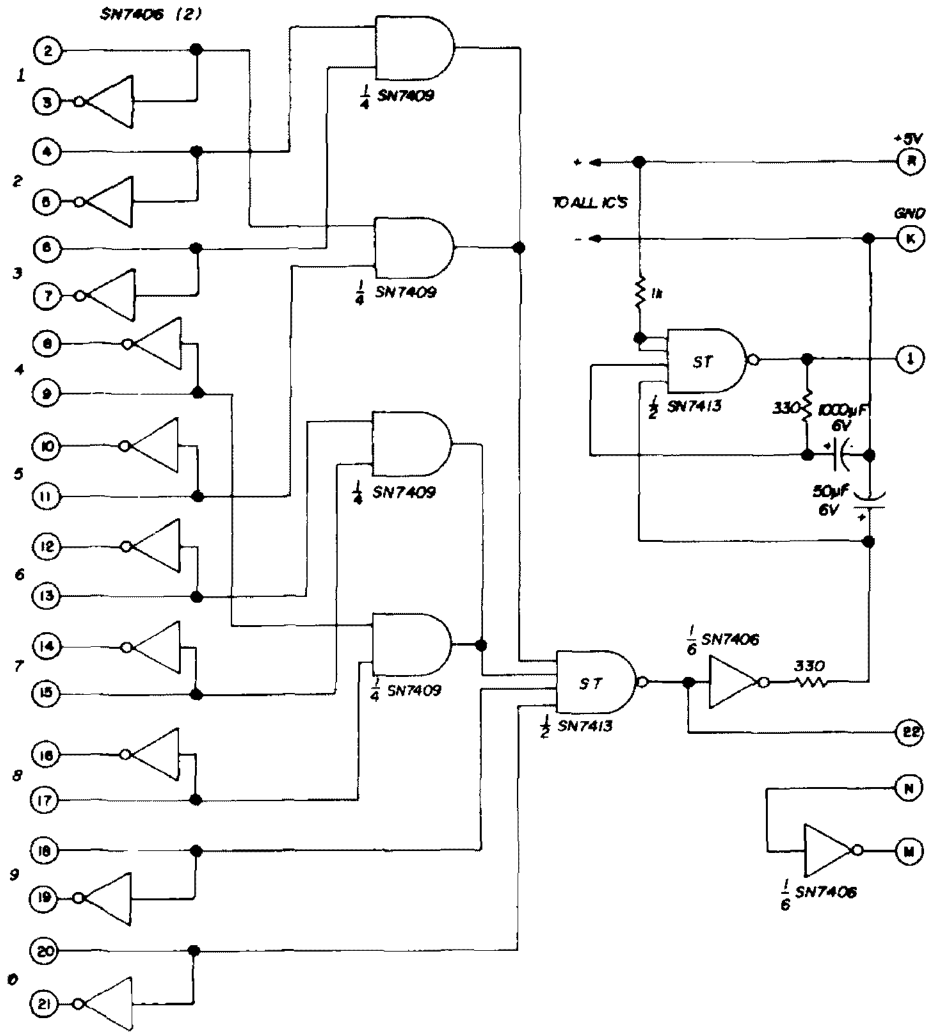


fig. 2. Diagram of the control module (referred to as circuit board number 1). Numbers in circles identify terminals on one side of board, letters those on other side.

feeding its output back to one of its inputs. The 330-ohm resistor and 1000-µF capacitor give a cycle time of about 0.63 seconds – 32 cycles in approximately 20 seconds. This is the clock pulse for the switching and timing circuits.

Its rate can be changed if desired by using a different value of capacitance. The 330-ohm resistor should not be changed as it will affect the operation of other parts of the circuit. Two unused inputs of the multivibrator are stabilized at high level by connection to V+ through a 1000-ohm resistor, as recommended by the manufacturer.

The fourth multivibrator input is connected to an inverter driven by the output of the other half of the SN7413, used here as a simple NAND gate. As long as this input to the multivibrator is in a high state, the multivibrator will oscillate under control of its own feedback circuit. When any Touch-Tone digit is received, however, it will cause a low input and high output at the NAND gate. The resulting low input to the multivibrator overrides the feedback circuitry, stopping the clock with its output in a high state until the tone signal ends. The 50-µF capacitor and its companion 330-ohm resistor introduce just enough delay into

Note: The control module and the buffer/driver printed circuit boards are two-sided which might make them impractical for home etching. The author can offer commercial-quality glass-epoxy boards for $10 for a set of three – the control module, the buffer/driver and one switching board of two modules. Additional switching module boards are $3 each. Boards are undrilled, but the price includes postage. Add $1 per board for drilling. The author can also supply the Texas Instruments ICs for those unable to get them locally. Write to R. B. Shreve, W8GRG, 2842 Winthrop Road, Shaker Heights, Ohio 44120.

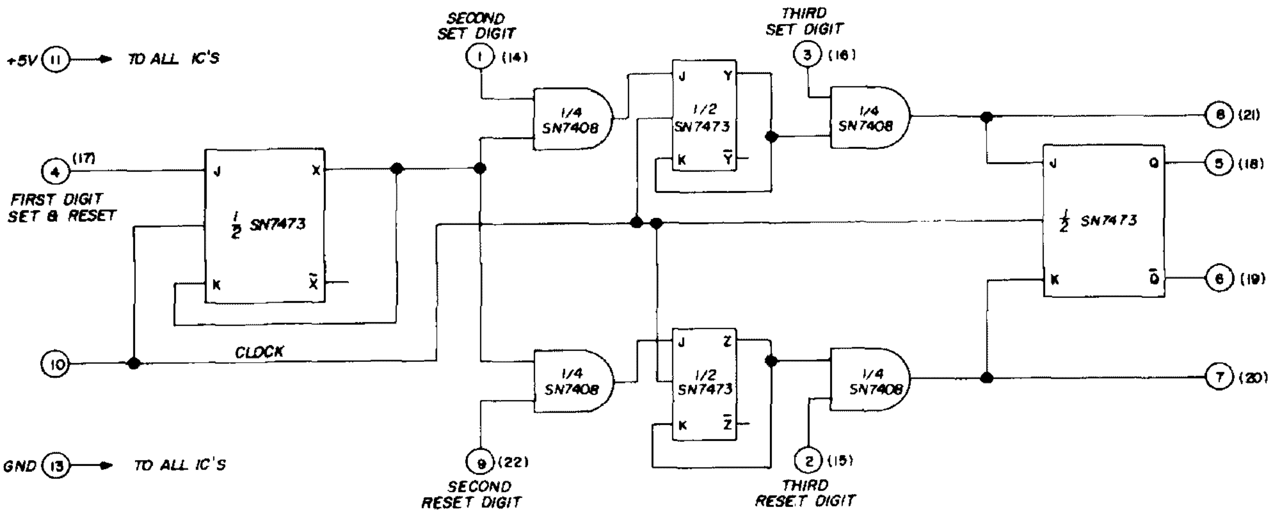


fig. 3. Basic three-digit sequential switching module. Each switching circuit board contains two identical modules and a common clock and power terminals. The terminals for one module are identified by the circled numbers, the terminals of the remaining module by the numbers in parentheses. These boards are referred to as circuit boards 3, 4 and 5. They are all identical.

the circuit so that the clock lags the Touch-Tone impulse by an interval which permits a switching module to recognize and act on the impulse during the clock cycle it generates.

Eleven of the twelve inverters in the two SN7406 ICs on the control module board are used in the control circuitry. Input and output of the twelfth are brought out to circuit-board terminals for use in an output circuit.

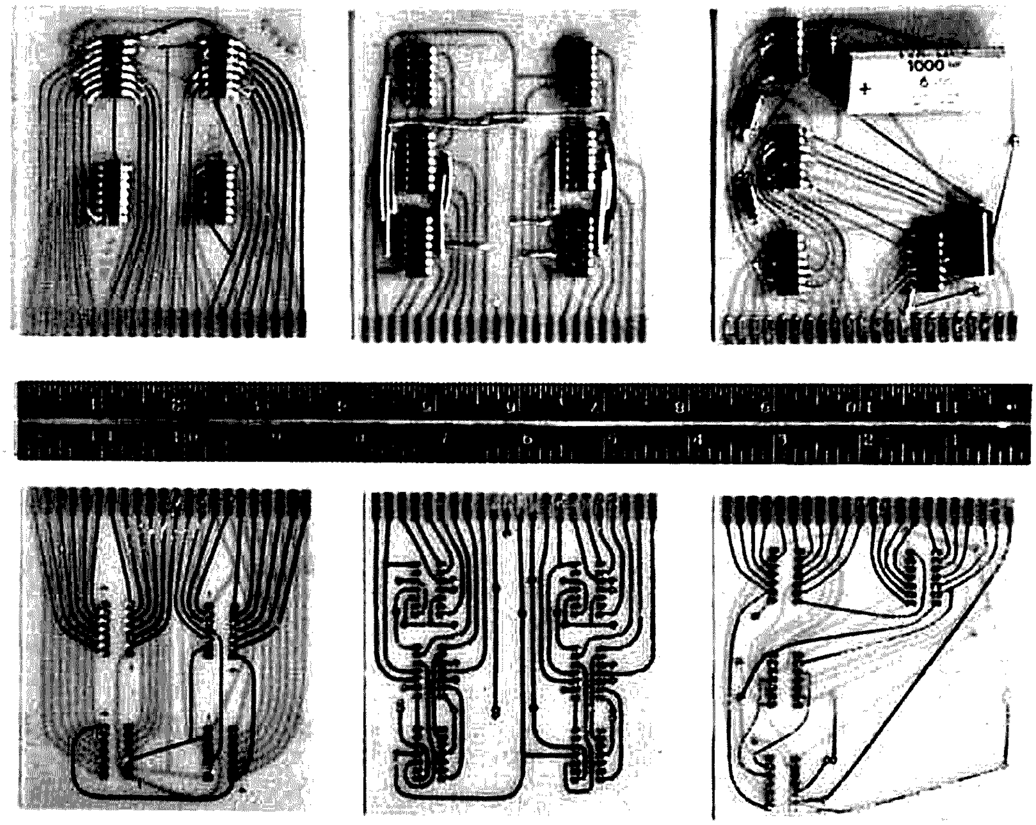
switching module

A typical three-digit sequential switching module is shown in **fig. 3.** It is made up of two SN7473 dual J-K flip-flops and a SN7408 quad two-input AND gate. If the clock input, terminal 10, is connected to terminal 1 of the control module, the switching module will react to the sequence of three digits connected to its *set* or *reset* terminals. There are two identical modules on each switching circuit board, with common clock, V+, and *ground* terminals. The remaining terminals for one module are identified by numbers in circles, and for the other by numbers in parentheses.

The *set* and *reset* codes for each module have the same first digit. For example, if 743 is selected as the *on* code for a function, the *off* code will also start with a 7. The module is programmed by connecting its *set* and *reset* inputs to appropriate inverter outputs on the control module. Assuming that 743 is the selected code, a Touch-Tone 7 will switch the first digit J input to the high state. This high is read into the flip-flop as the tone switches the clock to high, and the flip-flop will be *set* when the tone ends and the clock goes low.

The *set* flip-flop has a high X output for one clock cycle. If a 4 (the second digit in the sequence) is received during this clock cycle, it will act with the high X to generate a high J input at the second stage flip-flop. This flip-flop will in turn stay set for one cycle, during which presence of the correct third digit will set the final stage. The resulting high Q or low $\overline{Q}$ is used to control the desired function.

The third stage of the switching module does not reset automatically, since its K input is controlled by one of the *reset* sequence AND gates. It will switch back to its original state only when the proper *reset* sequence of digits is received. If a code sequence is interrupted or incomplete, the first two flip-flops of a switching module return to their standby state after being set for one clock cycle. False responses are minimized and no code addressed to one module will cause another to hang up even though they have one or two digits in common. The circuit will also distin-



Front and back views of the three circuit boards. On the left, the buffer/driver board; in the center a typical board with two switching modules; on the right the control module.

guish between sequences such as 743 and 734.

If on first thought a 0.63-second clock cycle seems too short an interval in which to send successive Touch-Tone digits, it should be remembered that the clock stops as long as a tone is held. Clock "low" time is also lengthened when a tone is received, as the 1000-μF timing capacitor is fully charged instead of being limited by the Schmitt-trigger hysteresis voltage.

The switching action has been described in simple form, with the switching-module clock input directly connected to the control board. In actual service, multiple switching modules require additional logic elements between the clock and the flip-flops. The SN7413 has a fanout capability of ten, and each switching module board has eight clock inputs, so SN7417 drivers are used to expand the clock output. One SN7417 will drive twelve flip-flops, so two are needed for each three switching-module circuit boards.

Additional control of the clock impulse is used to accomplish the objective, mentioned earlier, of having the repeater retransmit Touch-Tone signals intended for call-up or control of other stations without acting on them, and suppress transmission of signals used in its own control. A two-digit access code serves to identify the signals addressed to the repeater.

access module

The access module, board 6, is shown in fig. 4, which also shows related logic circuits on the other boards. The access module has two parts – a two-digit sequential switch similar to the three-digit switching modules, and a six-stage binary counter. The AND gate connection for access digit one is unnecessary, but the board was made by modifying a switching module board, and it was easier to connect the input to the gate than directly to the flip-flop.

In the access module, the control-module clock output drives the switching

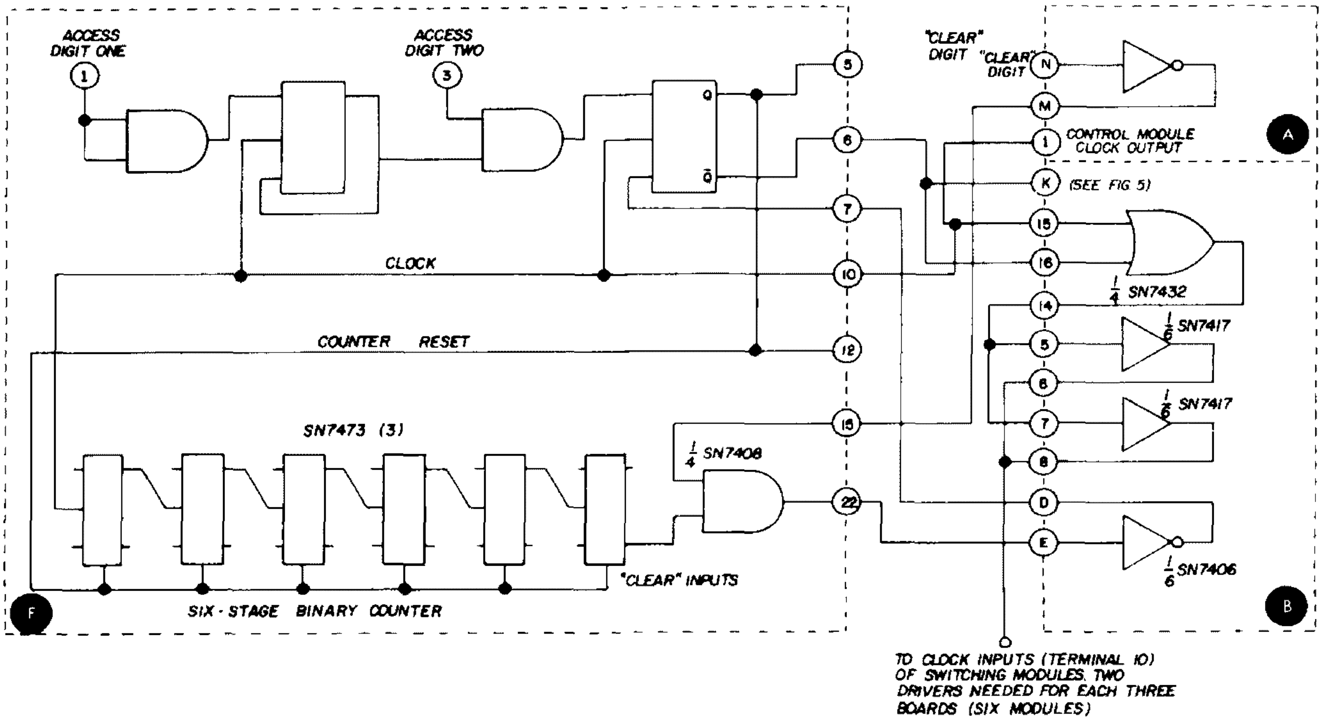


fig. 4. Diagram of the access module (circuit board 6) and related logic circuits on the other boards. The access module is labeled F, the control module is A and the output and driver logic on board 2 is B.

unit programmed to recognize the access code, and the first stage of the binary counter. It is also connected to one input of a SN7432 POSITIVE OR gate. The other input of this gate is driven by the $\overline{Q}$ output of the switching unit. As long as this output is high, the gate output is high regardless of the state of the control-module clock.

The gate actuates two SN7417 drivers, which in turn provide the clock impulses for all the switching modules on the other boards. Until the access switching module is set by receipt of the appropriate two-digit code, no clock pulses reach the switching modules on the other boards, so no Touch-Tone signals affect them. In addition, although the control clock is connected directly to the first stage of the binary counter, the counter clear inputs are connected to the Q output of the access switch, so it can not count as long as this switch output is low.

When the access switching module is set by the correct two-digit code, its Q output goes high and $\overline{Q}$ goes low. The SN7432 gate output now cycles with the control clock, activating the switching modules on boards 3, 4 and 5. The high Q output also releases the clear inputs of the binary counter, which starts counting the clock cycles.

As long as the access switch is set, the other switching modules will respond to the codes for which they are programmed. When the desired actions are completed, transmission of a single *clear* digit, generating a low at terminal 15 on board 6 resets the access switch, deactivating the other switching modules. If the unit is wired in this way, to clear with a single digit, that digit can not be used as part of an instruction code. If the *clear* input is not wired, or if the operator fails to transmit the *clear* signal, the binary counter will reset the access switch automatically after 32 clock cycles.

output logic

Design of the output logic circuits will depend on the functions to be performed and the nature of the equipment being controlled. Board 2 can be assembled with a variety of buffers, gates and drivers. Mine has a SN7406 inverter-driver, SN7417 non-inverting driver, SN7403 quad NAND gate and SN7432 quad POSITIVE OR gate. Use of some of this logic in the control unit has already been described. A good example of what

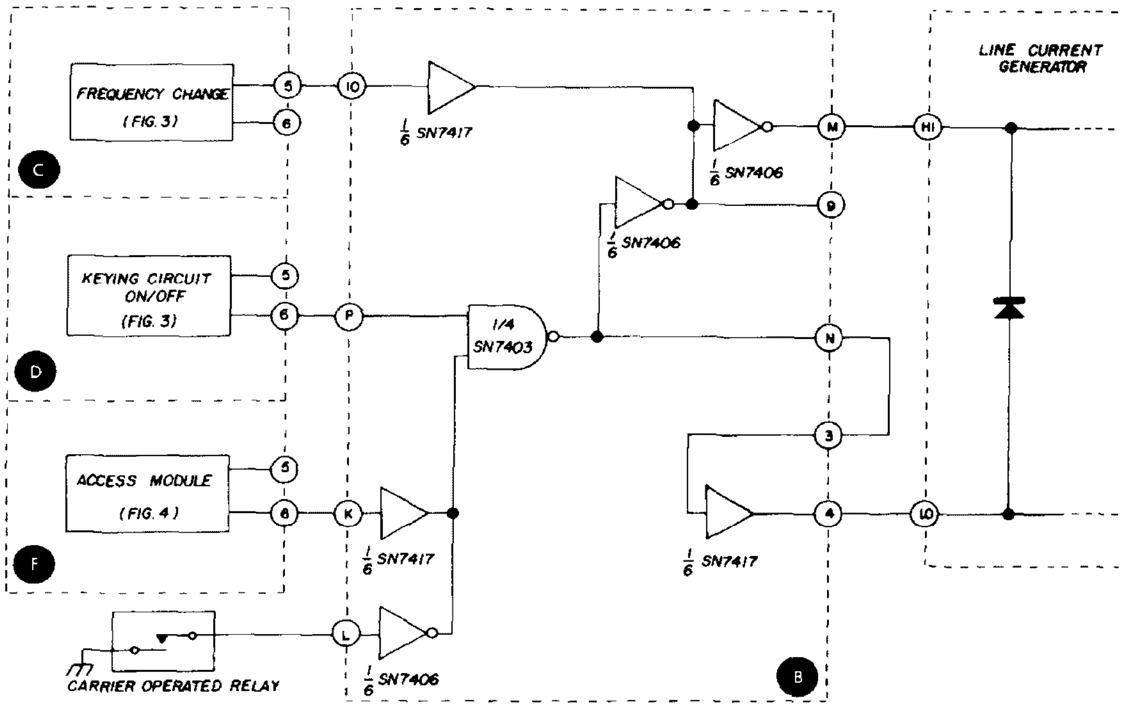


fig. 5. The keying control logic on circuit board 2 is labeled B. C and D are switching modules and F is the access module.

can be done by combining several outputs with appropriate logic is the keying control logic circuit shown in fig. 5.

As mentioned earlier, the receiver and transmitter are about a half mile apart, connected by a single telephone pair which carries both audio and dc keying current. The latter is supplied by a line current generator which will put either a low or high dc current on the line, depending on how it is keyed. The difference in current level may be used to select one of two transmitter frequencies. High line current flows when the generator *hi* input is grounded; low current flows if *lo* is grounded and *hi* is ungrounded. No current flows when both inputs are ungrounded.

For the *hi* input to be grounded, the input to the SN7406 inverter to which it is connected must be high. A low output from the *frequency change* switching module will prevent the *hi* input being keyed, while a high level output from it will permit the other part of the logic circuit to control.

Output of the SN7403 NAND gate is low when both its inputs are high. When low, it keys the *lo* generator input, and also keys the *hi* input if not inhibited by the *frequency change* switch. Low output of the *keying circuit on/off* switch prevents the transmitter being keyed. Likewise, a low $\overline{Q}$ output from the *access* switch, present when the switch is set to give signals access to the switching modules, prevents their retransmission.

If none of the logic is set to disable the keyer circuit, the carrier operated relay (COR) controls keying of the transmitter.

conclusion

Other control functions – silencing a single input, or switching the tone guard on or off, use very simple logic and will not be described in detail. It is hoped that what has been described will give you a picture of the unit's versatility.

Power requirements of the complete unit of six circuit boards, excluding the Touch-Tone decoder, total about 600 mA at 5V. Each additional pair of switching modules adds approximately 100 mA to the load.

reference

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ham radio

RTTY ribbon re-inkers

A quick conversion of a $2 surplus unit gives almost endless perfect page copy

Irvin M. Hoff, W6FFC, 12130 Foothill Lane, Los Altos Hills, California 94022

Any typing machine takes some kind of ribbon at periodic intervals in order to present a good appearance. Teleprinters for RTTY are no exception. Depending upon the amount they are used, a typical ribbon might last from a few weeks in a busy station to a few months in one less active. Cotton ribbons are normally used, as they are relatively inexpensive and can be tossed out when the page copy becomes too light to read easily. While nylon ribbons are much stronger, fray less readily and present a somewhat better appearance, they are also higher priced.

A few years ago I was given a ribbon re-inker. I managed to whack this unit up and mount it on my 28ASR, although it required drilling a hole in the case-hardened steel used on the model

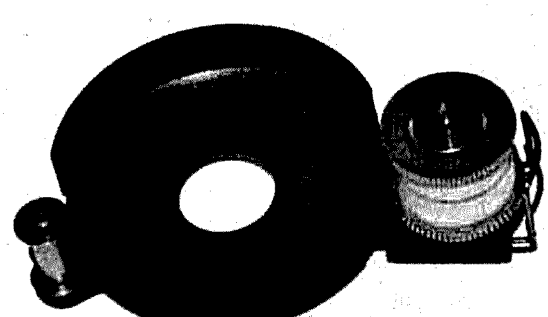


fig. 1. View of the basic reinker as used on the model 15 or 19. For the 28, this is inverted, the reservoir plugged into the other side and the unit modified as in fig. 2.

28 – no easy feat. A nylon ribbon came with the unit, and even though the teleprinter at W6FFC is running 7 to 10 hours per day, the same ribbon is still in use after some three years! Whenever it gets a bit light, I merely put a little additional ink in the reservoir, and the copy looks excellent again for a week or so more.

Normally, I do not write articles about hard-to-find surplus items. However, recently I was able to get a fairly large supply of these items and thus felt such an article might be of general interest. The kits were originally made for model 15 machines in the military services. The basic re-inker unit is shown in fig. 1 and has a detachable reservoir around which the ribbon is placed prior to its being wound on the spool. The top of this reservoir unscrews and ink is poured in, it then seeps out two small holes onto felt pads that surround the reservoir, and the ink then is distributed on the ribbon as it contacts the pads. One reservoir-full will take approximately one full ribbon traverse to distribute; once on the ribbon it then evens itself from layer to layer in a few hours or several traverses. The usual tendency is to overink, but after several years, I can usually get the results I want.

As it comes in fig. 1, the unit adapts immediately to the model 15 or 19 merely by removing one nut and exchanging this unit for the one already on these teleprinters. The re-inker also can be

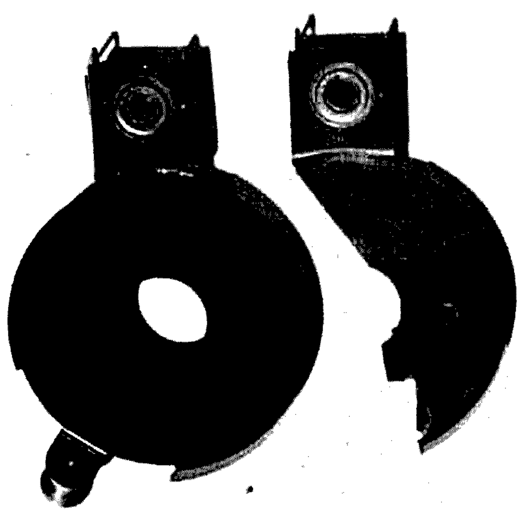


fig. 2 The reinker base: before and after.

modified quickly so it will go on a model 28 teleprinter.

Fig. 3 is a full-size template for cutting and drilling the re-inker. Cut out or copy the template, turn the re-inker so the flanged lip and post are down and place the template on the re-inker base with the legs of the template around the resevoir plate.

Mark lines B and C on the re-inker base and, using tin snips, a hack saw or a sabre saw, cut out at A (non-critical), B and C. The cut at C is critical and should be done carefully. Discard the shaded portions entirely.

Mark the hole, but do not drill it yet. Take the bolt out of the ribbon holder, and while holding the re-inker in position, swing it slightly until you can see your mark in the hole–check to make sure

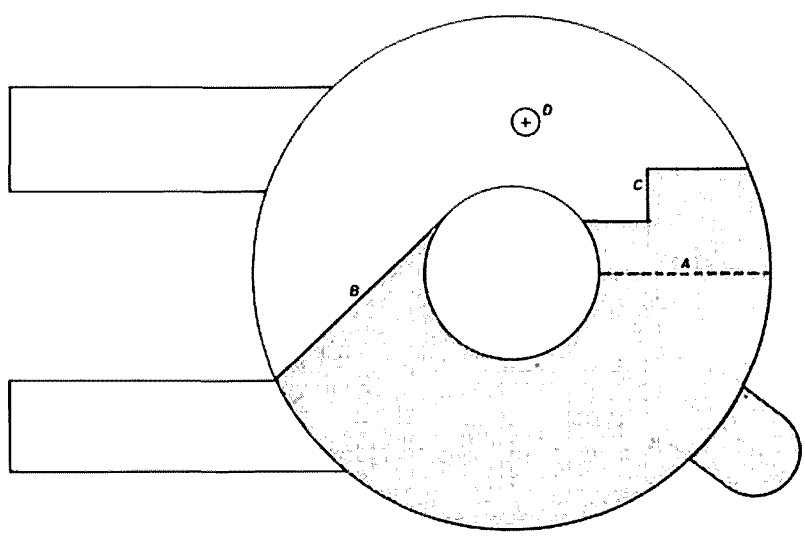


fig. 3. Template for modifying the basic unit to fit the model 28 or 35.

your hole is centered or remark it correctly. Drill the hole with a 7/64-inch drill, replace one short ¼-inch 4-40 bolt on the model 28 with a 3/8-inch 4-40 and fasten the re-inker in place with a nut. Beside changing the 4-40 bolt, there is no modification to the model 28 itself.

An exact-size template is furnished with each re-inker sent out by the author if a self-addressed stamped envelope is furnished for the purpose.

Possibly I got ahead of the story, so I will try to give a better explanation. On the model 28, the left-hand ribbon holder has a spindle on front for the ribbon to pass over, just to the right of this is a vertical gate through which the ribbon passes – when the ribbon comes to the end, a rivet in the ribbon catches in this gate, carrying it to the right, activating the reversing mechanism. If you remove the ribbon spool, you will note this reversing lever is pivoted at the rear and held in position with the ¼-inch 4-40 bolt previously mentioned. A collar prevents the lever from binding as the bolt is tightened. This is the bolt that is exchanged for a longer one, and it is to this bolt that the re-inker attaches. Fig. 4 shows how this appears when finished – it is not necessary to even remove the ribbon holder from the teleprinter. Fig. 5 shows how this appears from the bottom – and also shows how the ribbon goes through the gate, past the spindle and around the reservoir prior to being

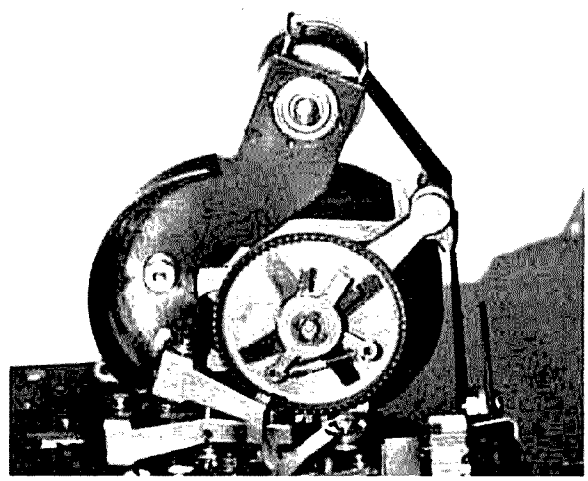


fig. 5. The underside of the ribbon holder with the reinker held in place.

wound on the spool. Fig. 6 illustrates the end result.

After getting the unit mounted, make sure the reservoir is at the same height as the ribbon spool itself – or of the spindle, and that it is vertical with respect to the spindle. Otherwise the ribbon might want to "overlay" onto itself a little as it goes past, indicating the reservoir is not centered properly. You can bend this up or down somewhat with a pair of needle-nosed pliers if needed, or with your fingers.

ribbon holder removal

Although not necessary, some of you may prefer to remove the ribbon holder entirely – it only takes a moment and does make the installation of the re-inker simpler, as well as offering a good means of determining if the reservoir is centered the best. There are two C-clamps and one spring to unhook, and the unit pulls right off the post around which it oscillates up and down. Fig. 7 shows the first retaining ring to be removed on the bottomside of the unit, the lever is then pulled free. Normally this lever causes the ribbon spool to oscillate up and down, allowing you to read the page copy when the typing unit is not striking a letter.

Fig. 8 shows the retaining ring at the top of the ribbon holder and fig. 9 shows the one spring on the right side of the

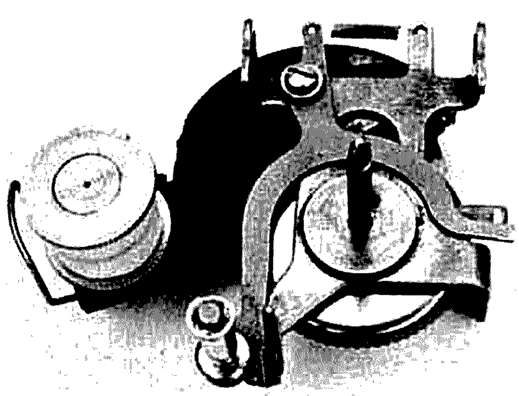


fig. 4. The ribbon holder has been removed to show how the modified reinker is attached for the 28 or 35.

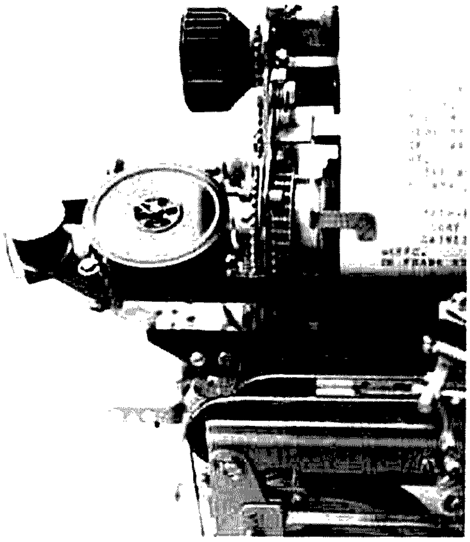


fig. 6. The complete reinker in use on a model 28 teleprinter.

ribbon unit as observed from the front. When reassembling the ribbon holder, refer to fig. 7 again and make certain the two levers shown on the left of the photo are interhooked as shown, which is simple enough to do while you are sliding the unit back into place.

other re-inkers

Other re-inker ideas have been tried and one is offered commercially. This latter system involves removing the ribbon from the machine, taking off the top plate of a special ribbon spool that comes with the kit, dropping ink at strategic places around the wound-up ribbon, and leaving it overnight for the ink to distribute itself uniformly. Then the special top is replaced and the ribbon and spool are placed back on the machine. This sytem never appealed to me.

Another system that was developed by Rex Peters, W4ZAG, merely was to take an empty ribbon spool and drill large holes in it at various places so that dropping some ink into the holes would allow the ribbon to disperse this ink over a period of time. Rex used this system for several years, but now uses the reservoir method described in this article. Other systems have been tried, but they seem to have little merit compared to the reservoir method.

ink

Various firms sell ink that is satisfactory. I use metal tubes of ink obtained from the National Cash Register Company – they resemble medium-sized toothpaste tubes. Ink used for ribbons is considerably thicker than ink used for fountain pens, so it is not as runny, although it is still in liquid form, and not really a paste. NCR ink comes in several colors; I use their black K-575. One tube is enough for at least a year's supply unless your station is quite active. Ink is available through any NCR distributor; look in the Yellow Pages in your phone book. The price has been 75c a tube.

When printed copy gets a little light, just remove the top of the reservoir and insert some ink – do not fill it too full or it will ooze out the hole in the top of the reservoir when it is replaced. You can put ink in regardless of the direction the ribbon is going, or whether it is at one end or in the middle. However, it will work the very best if you wait until the ribbon is all wound up on the right-hand spool and just starting to come back to the left. This puts ink on as it winds up, rather than as it unwinds. You can force the ribbon to go either direction you wish if you manually hold that reversing gate to one side for a few characters.

The usual tendency is to wait too long

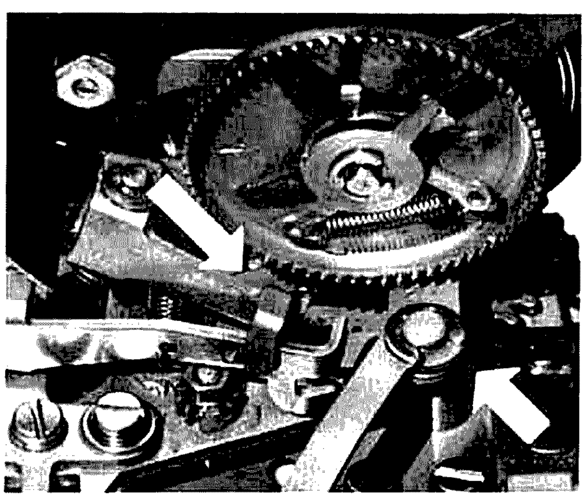


fig. 7. The bottom side showing the retaining ring (right) that is removed to take off the ribbon holder and (left) the manner in which the ratchet levers are unhooked.

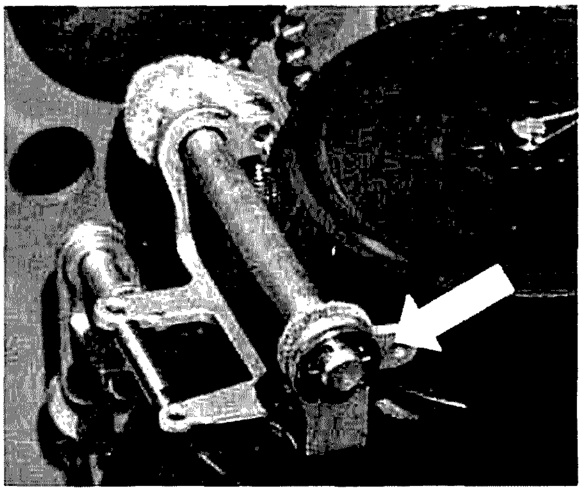


fig. 8. The retaining ring at the rear of the left spool is also removed to take off the ribbon holder.

and then over-ink. Do not put more than one reservoir full at a time on the ribbon. It may appear for awhile that nothing is happening and that it really needed lots of ink – you will want to put more on, but don't. Give it at least 24 hours or a few traverses of the ribbon (best of all, both!), and it will act like a new ribbon again. If you load several reservoirs full concurrently, you will soon find it is too black and certain characters will fill up and smudge, making terrible looking page copy for a few hours. This has been one reason a few people have not cared for re-inkers, but if you use some of these hints, you will not have the usual problems.

the re-inker kit

These kits contain the base plate, two reservoirs (one for a spare), a nylon ribbon, a large number of spare felts (after three years, I'm still using the original ribbon and original set of felts), a tool (plastic) for slipping the new felts onto the reservoir cylinder, a nice plastic box (excellent for storing radio parts), a small wire tool for cleaning out the seep holes in the cylinder (I've never needed to do this) and finally a plastic bottle of ink. This ink is pretty old, and too thick – I recommend you throw it out first thing. It will clog the seep holes and generally be unsatisfactory unless thinned – too much bother to mess with.

With this system, you can keep copy looking as though you had a new ribbon on, all the time, and still not get your fingers dirty. Even if you had unlimited access to free ribbons for merely carting them home, you would soon prefer using the same nylon ribbon with a re-inker.

The re-inker has been so nice I tried for some time (mentally) to see if I could figure out some method by which I could re-ink the nylon IBM cartridges on my Selectric typewriter (about $2.50 each). I gave up, although Dave, W7RSJ, says he has figured out a way to do it! Many typewriter ribbons could be re-inked on your teleprinter and then replaced on the typewriter, too!

availability

The author has a number of these available for $2, postage paid anywhere in the U. S. A. Please do not order from outside the U. S. A., as we have no facilities for going through customs, nor

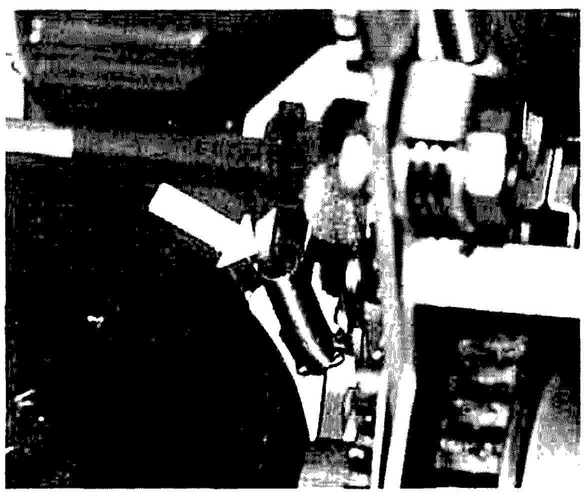


fig. 9. The third and final thing to remove the ribbon holder is to unhook this spring at the right of the ribbon holder.

an export license. Include an sase for an extra template for model 28s or to return your money in case we run out of the limited supply available. Send to:

Irvin M. Hoff W6FFC
12130 Foothill Lane
Los Altos Hills, Calif. 94022
Attn: Re-inker

ham radio

accurate noise-figure measurements for vhf

For the serious vhf experimenter a simple circuit and procedure for measuring vhf noise figures with accuracy

Louis N. Anciaux, WB6NMT, Box 1000, Dixon, California 95620

Building a low-noise preamplifier utilizing the solid-state devices available today, you can obtain very impressive receiver sensitivities at vhf.[1] For weak-signal work, such as meteor scatter, long-haul tropo and moonbounce, lowest noise performance obtainable is paramount. To ensure this, an accurate measure of the receiver's nf (noise figure) must be obtained. Adjusting for lowest noise performance using a noise source or a weak-signal source is unsatisfactory at best, and actual performance is unknown. The nf is still unknown.

Ready access to an automatic laboratory nf meter is generally not possible for many, so a suitable means of obtaining nf measurements must be found. Several nf meters utilizing the characteristics of a temperature-limited diode have been published previously.[2, 3] The Sylvania type 5722 is a diode specially constructed for use as a temperature-limited noise source. This tube, selling for approximately $8.40, exhibits predictable noise output response, which is readily determined by measuring the current passing through the diode.

noise factor

Using this diode, the *noise factor* of the receiver may be obtained when the noise output of the diode exactly

matches the noise output of the receiver. When this condition is satisfied, the *noise factor* is simply

$$20\,IR \qquad (1)$$

where I is the current thru the diode, R is the load resistance and the number 20 is a constant.[4] For the usual 50-ohm system, this reduces to 1000I; or the *noise factor* is equal to the magnitude of the current in mA passing through the diode. To obtain *noise figure* (nf), merely take ten times the logarithm of the *noise factor*.

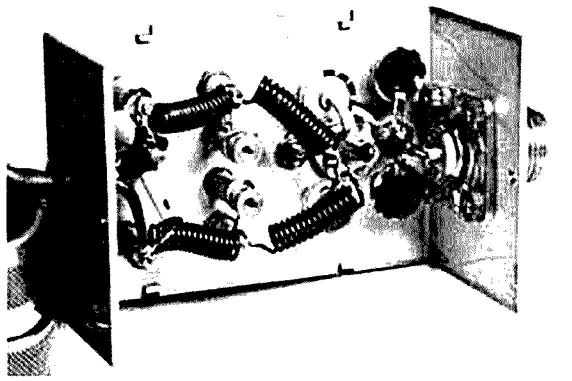
design notes

To make accurate nf measurements, you must know the load resistance and current through the meter with a fair degree of certainty. Previous models have failed to consider the load resistance as an rf network, using only a 50-ohm resistor and excessive lead lengths. Above 100 MHz especially, the small reactances associated with lead lengths of the order of 0.5 inch are sufficient to cause an error approaching 1 dB when measuring receivers with under 3-dB nf. Current through the diode must be known with precision, 5% error being an excessive amount.

The nf meter described here gives excellent performance and repeatable results through the 432-MHz band. Comparison with industrial meters of high accuracy have shown excellent agreement. Error is under 0.1 dB for the bands 220 MHz and below, and at 432 MHz, the meter consistently reads within 0.5 dB of the industrial meter. At 432 MHz, W6FZJ consistently measured preamps at 0.3-dB higher with the meter described here than with a commercial meter.

precautions

Be careful with the rf head construction and with the meter circuitry. A large scale meter is a must, as the current must be measurable to less than 0.1 mA. Since most nf measurements of greatest interest are under 5 dB, a 4-inch or larger scale meter with full-scale deflection of 5 mA is an excellent choice. You can use a shunt to obtain readings up to 25 mA, as desired. If you use a meter with less than a 5-mA movement and a shunt resistor, it must be calibrated against a known standard. Particularly under 3 mA, readings *must be accurate* to less than 0.1 mA. Note that the current path is directly



Interior view of the rf head. Readily visible are the various components and their placement within the minibox. The three 150-ohm resistors must be placed to ensure shortest possible leads. Low-loss ceramic socket for the 5722 keeps losses low at higher vhf ranges.

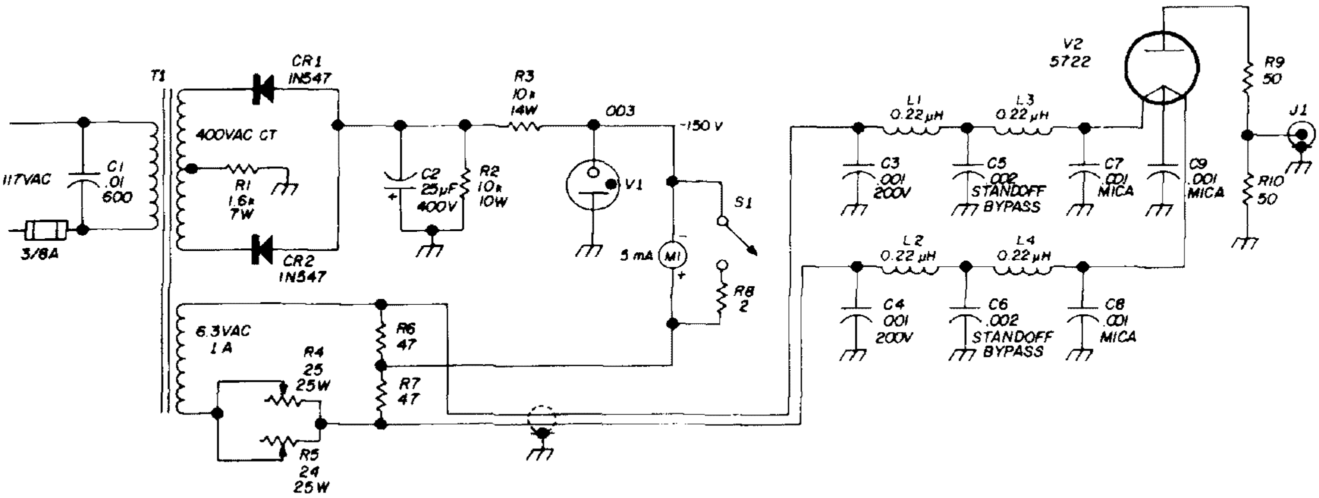
through the meter and should the rf head develop a short, it means an instant demise for the meter. You may add some sort of meter protection if you want.

construction

To begin building the rf head, place the tube socket near one end of the minibox, so that R9, the 50-ohm quarter-watt resistor, will just fit between the anode pin and the center pin of the connector which is mounted on the end of the minibox. Either three 150- or four 200-ohm quarter-watt resistors may be used for the load, R10. Carefully select these resistors to obtain a near match between them, as well as a final paralleled value as close to 50 ohms as possible. By placing them symmetrically around the UG-58A/U connector, their leads, soldered directly to the connector body and to the center pin, are no longer than 1/32 inch. This will ensure a load resist-

ance which exhibits a very few ohms reactance at 432 MHz.

The single 50-ohm quarter-watt resistor from the anode pin to the connector center pin acts as a swamp to minimize swr effects from the receiver non-critical, as long as the voltage does not vary radically and is free from excessive ripple. A 100- to 200-volt supply is satisfactory, and will be beyond the diode's space-charge potential sufficiently to ensure repeatable results.[5] The VR



J1	UG-58A/U chassis jack	R10	three 150-ohm ¼W or four 200-ohm ¼W resistors selected to yield 50 ohms when paralleled
L1,L2 L3,L4	approx. 0.3 µH, 16 turns no. 20 wire, ¼-20 bolt used as mandrel. L1 and L2 may be up to 10 µH.		

fig. 1. Complete schematic of the noise figure meter including the power supply and the rf head. They are shown here connected with shielded cable. Consult the text for modification to the inductors for different frequency ranges.

input circuitry which can cause erratic readings. The uhf button-mica bypass capacitors at each of the three cathode leads should be located to keep lead lengths as short as possible, and of equal length. The rfcs farthest from the 5722 socket may be made larger, up to 10 µH, to allow use well below 100 MHz. The model shown here is used mainly for 220-MHz preamps; thus both rfcs were chosen as 0.3 µH.

power supply

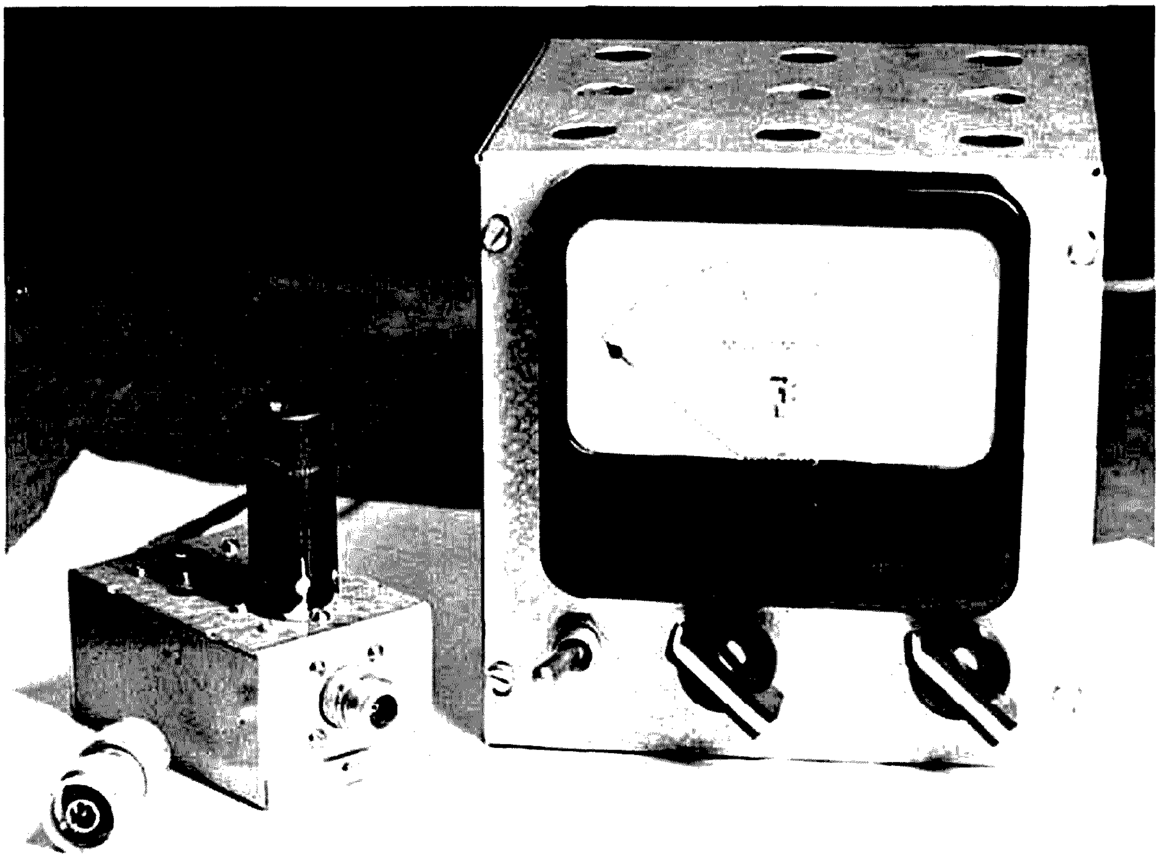
The best way to control filament voltage is with a small Variac in the primary of a separate 6.3 V transformer. A second method, as shown, uses two power rheostats in parallel. One allows the filament voltage to be increased to the point of diode conduction (approximately 1 mA) and the other acts as a fine-tuning control to set the current to the appropriate level. The dc supply is tube is not necessary, as only the current through the diode affects the noise produced, and small voltage variations will not hurt.

operation

To use the meter for *determining* nf, you must be able to accurately measure a 3 dB change in noise output of the receiver. When the noise output of the 5722 diode equals that being produced by the receiver, the nf of the receiver may be determined from the diode's characteristics. A vtvm, or a vom on low-ac range, may be used to measure the receiver's audio output. In this case, the receiver must use a product detector and the agc must be disabled. In receivers using square-law detectors (diode types), you must use the 3-dB pad insertion method. Some receivers have very responsive S-meters which may be used as indicators when using the 3-dB pad.

The 3-dB pad insertion method is the most accurate, and therefore, the preferred method of obtaining nf measurements. Connect a speaker to the receiver output and set the volume to some reference level. Attach the nf meter rf head directly to the converter or preamp input. Insert a 3-dB pad into the i-f line between the converter and receiver, noting a decrease in output from the receiver. As you turn on the 5722's filaments slowly, and as the diode current increases, the noise output of the receiver will also increase. When the output of the receiver is the same as the reference level prior to insertion of the pad, read the diode current and turn the filaments off.

Removing the pad should return the indicator to its original reference level. If it does, then repeat the procedure to obtain another diode-current reading. If the diode current is the same, then change the reference level to another value, and repeat the same procedure to obtain diode current. If all diode-current readings are the same and are repeatable, then it is generally safe to assume the readings are valid. However, if the readings vary from trial to trial, then precautions are necessary.



The noise figure meter and rf head. The 4½-inch scale meter allows direct reading to 0.1 mA with interpolation to 0.025 mA readily possible. The interconnecting cable allows the rf head to be used separated from the nf meter, facilitating measurements. The UG-57B/U adaptor connects two UG-58A/U chassis jacks together directly, keeping losses to a minimum.

problems

If you obtained erratic diode-current readings, there are several possible causes. First, if the audio output is being read on a vtvm and a 3-dB change is sought, inserting a 3-dB pad in the i-f line should also indicate a 3-dB change on the vtvm. If it does not, then either the agc is not disabled or you are overloading the mixers or detectors. Second, if the swr between the converter and receiver is not 1:1, then using two additional pads should cure this problem. Inserting the

3-dB pad between these two pads will allow you to make repeatable measurements. The two additional pads act as swamps to stabilize the circuitry. Third, if the converter or preamp-input circuit is grossly mis-adjusted, (not 50 ohms) erratic readings will also occur. Fourth, if the converter or preamp is not properly neutralized, very erratic readings result.

If a preamp, previously measured into another converter at a low noise figure, shows considerably higher nf, it may be due to very high nf in the second stage (converter input). This is seen from the relationship:

$$F_T = F1 + \frac{F2}{G1} \qquad (2)$$

where F1 and F2 are *noise factors* of the first and second stages, F_T is system *noise factor,* and G1 is the *gain factor* of the first stage.[4] In this case, the second stage nf must be reduced to an acceptable level, either by redesigning or by adding another rf stage.

For example, consider a 2-dB nf preamp with 10-dB gain and a converter with a 15-dB nf. Converting dB to *factors,* 2 dB = 3.01, 15 dB = 17.61, 10 dB = 10; and plugging into the equation yields a *noise factor* of 4.77 and a system nf of 3.0 dB. This shows the importance of having the second stage nf reasonably low in order to enjoy fully a good preamp's performance. In the case of a 5-dB second stage, system nf would be 2.34 dB. Increasing the gain of the preamp to 15 dB would further reduce the system nf to 2.2 dB.

procedure

As an example in making a nf measurement, connect the rf head of the nf meter to your preamp input. The converter output is fed into a 6-dB pad, through a piece of coax to another 6-dB pad and into the i-f receiver. The S meter on the receiver reads S-3. Inserting the 3-dB pad in the i-f line after the first 6-dB pad, shows a drop to nearly S-2. The diode filament current is slowly increased until the noise produced by the diode just equals the S-3 reference level. The diode current is read as 2.3 mA. The diode filaments are turned off and the 3-dB pad removed from the i-f line. The S-meter should again read S-3.

Repeat the above procedure and obtain another diode current. Assuming it is again 2.3 mA, remove one of the 6-dB pads and insert a 10-dB pad in its place, obtaining a new reference level. The diode current obtained now should be the same if all is well, otherwise, problems exist. Make at least two tests to obtain diode current.

To convert this reading of 2.3 mA to nf, simply take ten times the logarithm of 2.3. That is, log 2.3 = 0.362, thus nf = 3.62 dB. The accuracy of the current meter becomes apparent at this point. If the accuracy is only ± 0.1 mA, then the current may be 2.2 mA or 2.4 mA. The nf corresponding to these currents are 3.4 dB and 3.8 dB. The nf of the receiving system is 3.6 ± 0.2 dB. When making readings around 1.5 mA, ± 0.1 mA error corresponds to nf of 1.46 dB at the low- and 2.04 dB at the high-error point. This 0.5 dB or so difference in nf at this level represents more like 1 or 2 dB in receiving sensitivity, and for moonbounce work, those dB really count.

I would like to express my appreciation to W6FZJ for originally bringing to me the idea of the distributed impedance load in the rf head, and to W7CNK for many enlightening discussions during which we both came to understand many of the reasons for doing things the way we have and for his helpful criticism in reading the drafts of this article.

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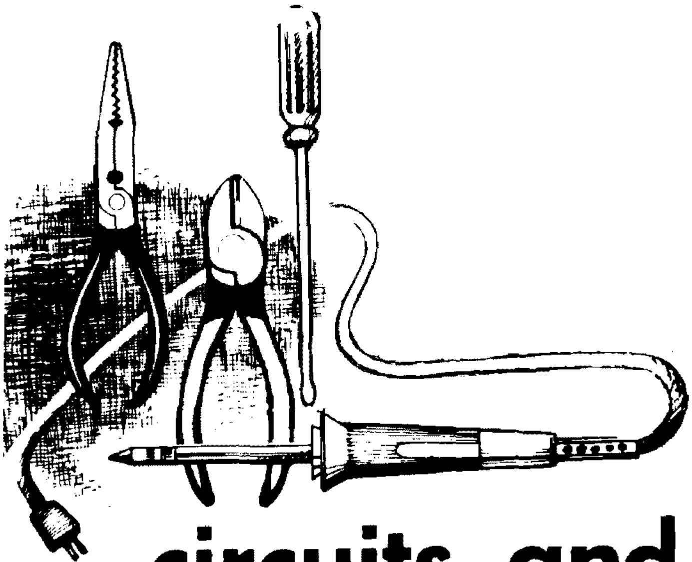


circuits and techniques

ed noll, W3FQJ

digital multivibrators

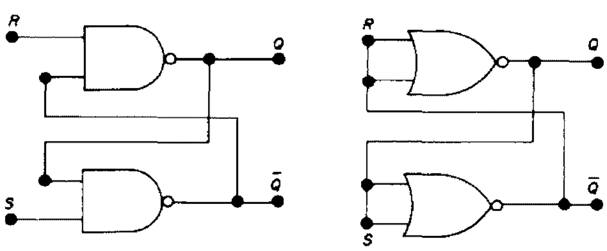
***Old-fashioned circuits** acquire new names.* One of the most common electronic circuits is the bistable multivibrator which, for more than three decades, has been standard in television and radar circuits. It is often called a toggle or flip-flop multivibrator. In the clipped computer language it is called an FF. There are a variety of multivibrator types to be found packaged in digital integrated circuits. Although some refinements have been made by the computer industry, the multivibrator circuitry is rather fundamental as compared to some of the earlier versions used in radar and television equipment. You old timers will remember the phanastron!

Two vacuum tubes or two transistors fed back upon themselves can be made to flip-flop. A very basic multivibrator can be formed by feeding back two NAND or NOR gates upon themselves as shown in fig. 1. In computer vernacular the inputs are called *set* and *reset* or simply S and R, while the outputs are designated Q and Q. These correspond to the trigger inputs and the MV outputs of the past. The block symbology and the truth tables are also shown in fig. 1.

The Q and $\bar{Q}$ can be given logic 0 and logic 1 designations. The *set* or logic 1 state of such a flip-flop corresponds to a Q output of 1 and a $\bar{Q}$ output of 0. The converse of the above (Q of 0 and $\bar{Q}$ of 1) is spoken of as the *reset*, cleared or zero-state position. Customarily you say that the circuit has been *set to 1* or *reset to 0* representing the two states of the *flip-flop. These designations are shown on* the truth table.

When there are simultaneous logic **1** inputs the next state becomes indeterminate and in truth chart language it is said to be not allowed, forbidden or indeterminate. In general, simultaneous logic 0 inputs represent a no-change condition and values are related to the previously set or reset position.

An additional input can be added to a multivibrator which permits toggling or flip-flop operation. This input applies so-called *clock* information to both inputs simultaneously, fig. **2A**. As per the fundamental triggered-multivibrator, the output frequencies are one-half the clock



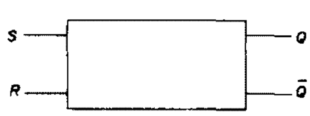
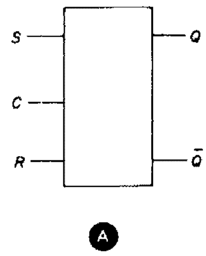


fig. 1. The basic flip-flop multivibrator and its truth chart.

S	R	Q	$\bar{Q}$
1	0	1	0
0	1	0	1
1	1	——————— forbidden	
0	0	——————— no change	

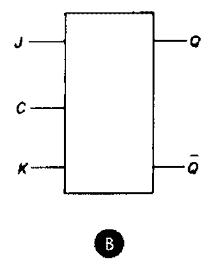
input frequency. This is the basic binary multivibrator, or 2-to-1 divider, or 2-to-1 counter. The logic of the output is controlled by the set and reset inputs of the clocked R-S multivibrator.

A popular multivibrator circuit included in digital ICs is the J-K flip-flop, **fig. 2B**. This arrangement permits ease in setting the flip-flop to a desired condition. In addition to the usual set and reset positions as shown in the truth chart, it is possible to supply logic 1 level to the J and K inputs simultaneously. If the clock pulse is also of logic 1 level there will be a switchover of the output logic voltages. However, if both the J and K inputs are logic 0 the clock pulse is not able to change over the output logic.



S	R	Q	$\bar{Q}$
1	1	— — — — — — forbidden	
0	1	0	1
1	0	1	0
0	0	— — — — — — no change	

fig. 2. Clocked R-S and J-K flip-flops.



J	K	C	Q	$\bar{Q}$
0	1	1	1	0
1	0	1	0	1
1	1	1	— — — — complement	
0	0	0	— — — — no change	

all other combinations — no change

A popular digital multivibrator is the universal flip-flop such as the 7472N shown in **fig. 3**. This combination permits you to set up a variety of multivibrator functions because it has both input pairs, R-S and J-K. The use of the J, K and C inputs permit J-K FF operation. Since the R and S inputs are overriding, the multivibrator can also be operated in the R-S manner. For simple operation as a 2-to-1 divider, you need only apply signal to the clock C input.

experiments

You can learn much by experimenting with several of the low-cost digital ICs.

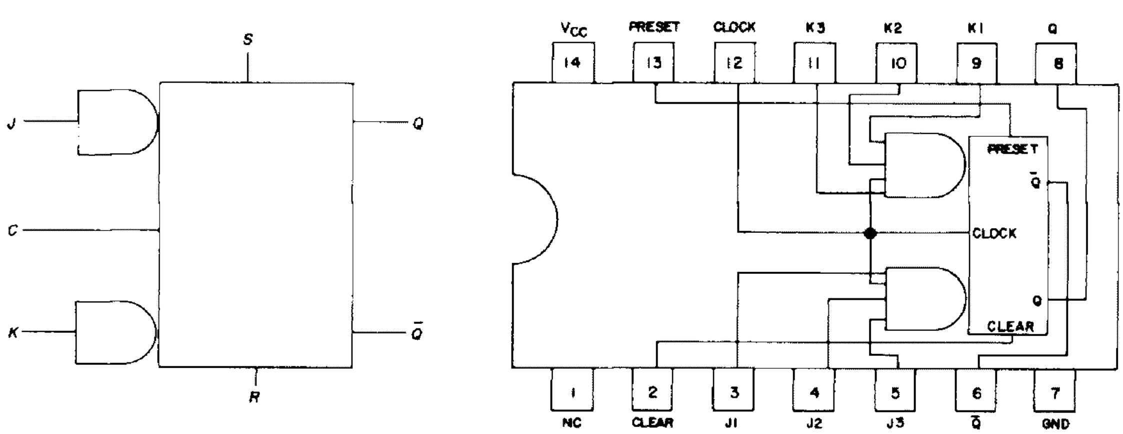


fig. 3. Universal flip-flop (A) and the actual pin connections of the 7472N (B).

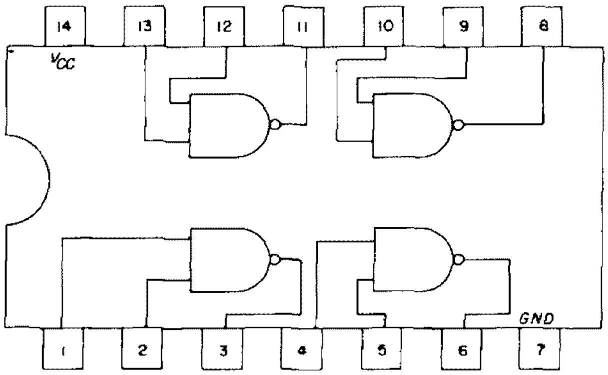


fig. 4. Connection diagram for the 7400 quadruple 2-input NAND gate.

Later these can be combined into a calibrator, counter or some other unit that can be useful in your station.

A small perforated board, 4½ x 5½ inches, supported on four stick-on protector pads can serve as a versatile test board. Several 14-pin in-line sockets can be attached to the board. All socket terminals, or as many as are needed, can be wired to soldering lugs that are screwed to the board. This arrangement gives you versatility and better experimental wear-and-tear with frequent rewiring when various types of digital ICs are to be used.

experiment 1
NAND gate

A readily available and low-cost digital IC NAND gate is the SN7400N (7400 series) which consists of four individual NAND gates with dual inputs, **fig. 4.** Supply voltage is connected to pin 14 for this TTL device, while common (ground), is connected to pin 7.

Connect the circuit of **fig. 5.** Use just one of the gates in the device. The circuit connections are simple; the dual inputs

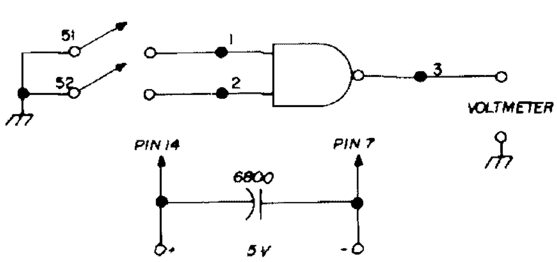


fig. 5. One-quarter of an SN7400N NAND gate connected to show dc operation.

are pins 1 and 2 and the output, pin 3. Connect a voltmeter across the output. Two spst switches permit you to establish the logic 0 and logic 1 inputs. This device has two states as in any binary logic system. The logic 1 state or *high* is represented by a voltage greater than plus 2 volts (positive logic) while the logic 0 state or *low* is represented by a voltage of 0.8 volts positive or less.

When the gates are open, the emitter voltage in the basic TTL circuit of **fig. 6** is more than 2 volts positive, and the emitter junction is being reverse biased. Open both switches and measure the dc voltage at pins 1 and 2 to confirm this statement.

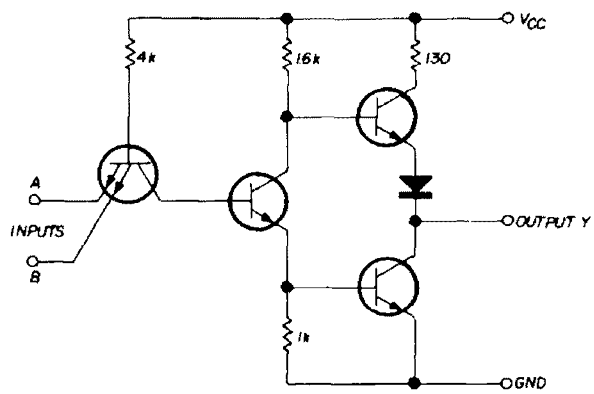


fig. 6. Basic circuit for a single gate in a 7400 digital IC.

The collector junction is forward biased and the collector current present in the base-emitter circuits of the output transistor results in a saturation current and 0 output voltage as indicated by the voltmeter. In this connection then, logic 1 voltage is present at pins 1 and 2 while logic 0 voltage appears at the output.

Now close switch S1. What happens to the voltmeter reading? Measure the voltages at pins 1 and 2. Note in particular that the voltage at pin 1 represents logic 0. Logic 1 voltage remains at pin 2 and the output voltage is also logic 1. When switch S1 is closed the input emitter junction is forward biased and the collector current now changes direction and cuts off the output transistor. Its col-

lector voltage rises to near the supply voltage and produces logic 1 output.

Repeat this step with switch S2 closed and S1 open. Repeat again with both switches S1 and S2 closed. Set down the truth table and compare it with the truth table of a NAND circuit as given previously in the April column.

experiment 2
flip-flop

Rewire the experiment board to the circuit of **fig. 7**. Two of the NAND gates of the device are now wired together. The output at pin 3 is connected to pin 4 while output 6 is fed back to pin 2. This

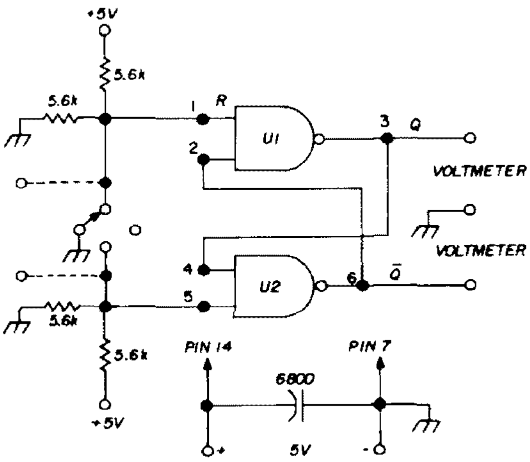


fig. 7. NAND gates in flip-flop. U1 and U2 are each ¼ of an SN7400N. Dashed lines show connections for an audio input.

feedback path is normal for a flip-flop multivibrator.

Connect the voltmeter between output 3 and common. Connect ground to input 1 with the switch. Is output 3 logic 1 or logic 0? Connect the voltmeter to output 6. Is it logic 1 or logic 0?

Restore the voltmeter to output 3. Set the switch to its unused center position. What happens to output 3 and output 6?

Connect the voltmeter to output 3. Use the switch to ground input 5. What happened as the switch closed? Check the outputs at 3 and 6. Note how the output is toggled as the switch is changed from input 1 to input 5 and vice versa. Complete the truth chart of **fig. 8**.

1 R	5 S	3 Q	6 $\bar{Q}$

fig. 8. NAND gate flip-flop truth chart to be filled in from actual experimental readings taken with the circuit of fig. 7.

experiment 3
ac flip-flop

Change the input of the flip-flop, **fig. 7**, in accordance with the dashed lines. Set the switch to its unused center position. Apply a 1000-Hz sine wave to the input. Connect the oscilloscope between output 3 and common.

Slowly increase the amplitude of the audio signal. When the audio level is adequate the sine wave will toggle the flip-flop. Note the quality of the output square wave. It has steep leading and trailing edges. Flip-flops are often used to sharpen the edges of slower rising and distorted waveforms.

Increase the level of the audio input wave. Observe how constant in voltage the output remains once the input wave is sufficient in magnitude to cause toggling. Try audio input frequencies of 100, 10,000 and 100,000 Hz. The quality of the output waveform remains good and the amplitude constant.

Reset to 1000 Hz. Display four square

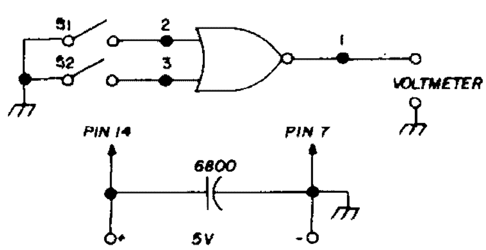


fig. 9. A 7402 NOR gate set up for dc operation.

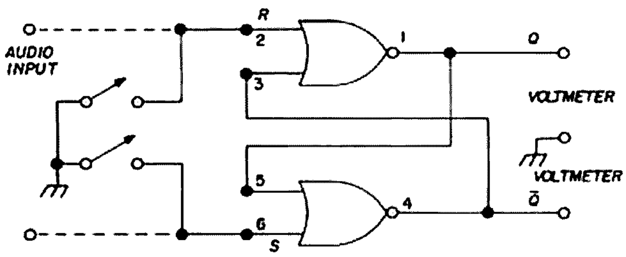
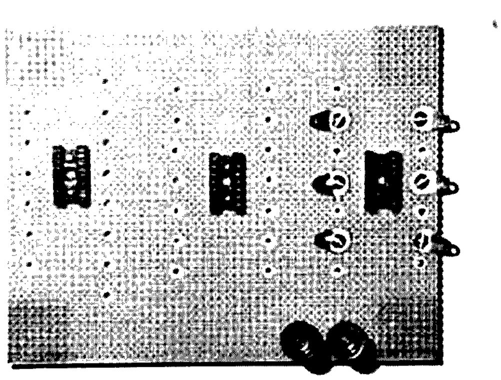


fig. 10. NOR gates arranged into a flip-flop.

waves on the oscilloscope screen. Transfer the oscilloscope to input 1 and note that output frequency and input frequency are the same.

experiment 4
NOR gate

Remove the 7400 positive NAND gate and prepare to use a 7402 NOR gate IC. Connect the circuit of fig. 9. Positive voltage is applied to pin 14 and ground to pin 7 as for the 7400. In this case, however, pins 2 and 3 are the inputs while pin 1 is the output. Switching arrangement is the same as that of the NAND circuit shown in fig. 5. Open is again logic 1; the switch closed to common corresponds to logic 0. Connect the voltmeter to the output and set down the voltage readings with both switches open, both switches closed, S1 open and S2 closed and, finally, S2 open and S1 closed. Prepare your own truth chart and compare with the NOR logic truth table of April's column.



W3FQJ's breadboard for IC experiments uses IC sockets and solderless binding posts for quick circuit changes while still protecting the delicate IC wire leads.

Open both switches and apply a 1000-Hz tone to the input. Connect the oscilloscope across the output. Remember, to activate the NOR gate both inputs must be driven to the logic 0 state. Therefore, when driven from a single audio source, the two inputs must be paralleled.

Slowly increase the audio level. After passing a certain point you will obtain NOR gate output. Again, a good square wave is obtained. Check the output on 100, 10,000 and 100,000 Hz.

2 R	6 S	1 Q	2 $\bar{Q}$

fig. 11. NOR gate flip-flop truth chart to be filled in from actual experimental readings taken with the circuit of fig. 10.

experiment 5
NOR gate flip-flop

Connect the flip-flop circuit of fig. 10. To observe the operation of the NOR gate flip-flop you will need two spst switches. Use the voltmeter to measure the voltage at outputs 1 and 4. Record these outputs for switches S1 and S2 closed, S1 and S2 open, S1 open and S2 closed, and, finally, S2 open and S1 closed. Note that for both S1 and S2 closed, logic 0, the output is indeterminate and is a function of whether S1 or S2 was closed previously. Complete the truth chart of fig. 11.

experiment 6
master crystal oscillator

A simple dual-input NOR gate, fig. 12, can be connected externally into a feed-

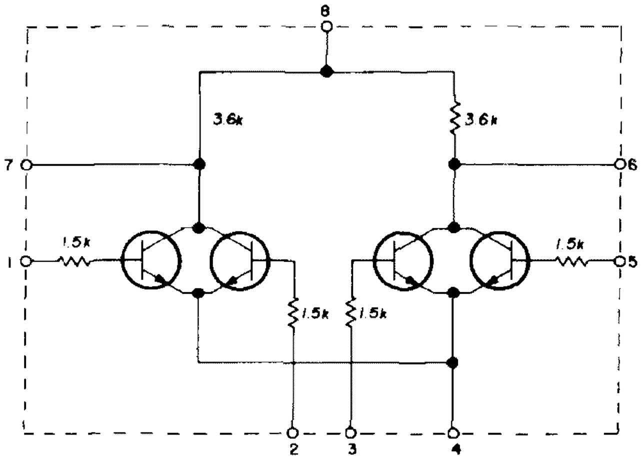


fig. 12. HEP580 MRTL dual-input NOR gate.

back circuit to perform as either a free-running multivibrator or a crystal-controlled square-wave generator. In the latter case, a crystal is inserted in the feedback path between output 6 and paralleled inputs 1 and 2. The second feedback path is provided by a capacitor inserted between output 7 and inputs 3 and 5.

The HEP 580 IC has an 8-pin base. Mount an 8-pin round socket on the breadboard. Assemble the circuit shown in **fig. 13**. Only four external parts are required: two resistors, a capacitor and the crystal socket. This is one of the simplest 100-kHz crystal calibrators you can make.

The square wave output can be observed at pins 7 and 6. The harmonic content of the IC output is high and clear

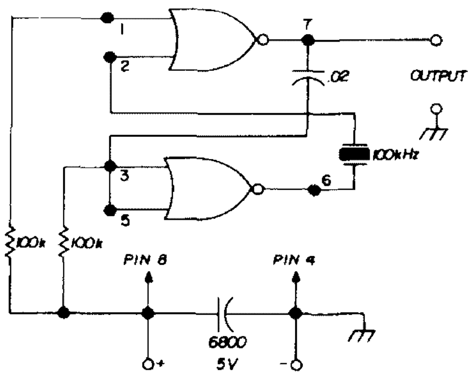


fig. 13. 100-kHz crystal oscillator using the HEP580.

calibration points can be found on frequencies as high as 52 MHz.

experiment 7

2-to-1 flip-flop counter

The addition of a low-cost J-K flip-flop multivibrator to the output of the 100-kHz generator provides you with a 50-kHz output as well. A TTL SN7472 will do, **fig. 14**. It can be inserted into the same in-line socket used previously in experimenting with the NAND and NOR gates. Supply voltage and common are again attached to pins 14 and 7 respectively. The output of the 100-kHz generator, pin 7, is applied to the clock input, pin 12. It is possible to obtain 50-kHz output at either the Q or $\overline{Q}$, pins 8 and 6

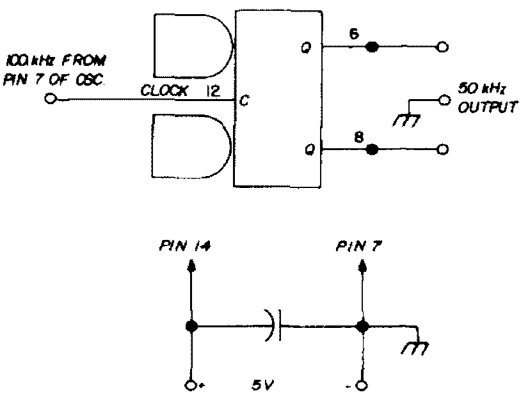


fig. 14. An SN7472 J-K flip-flop used as a 2-to-1 divider.

respectively, of the counter.

Observe the input signal at pin 12 with the oscilloscope. Display four cycles of the square wave. Now transfer the oscilloscope to the Q output, pin 8, and note that the frequency has been halved. A 2-to-1 count has been made.

Calibration points can be heard readily up into the six-meter band. In using a calibrator, the oscillator alone (divider switched off) can be used to locate 100-kHz points. Output of the divider can then be used to locate the 50-kHz intermediate points.

Next month additional counter stages will be added to the calibrator.

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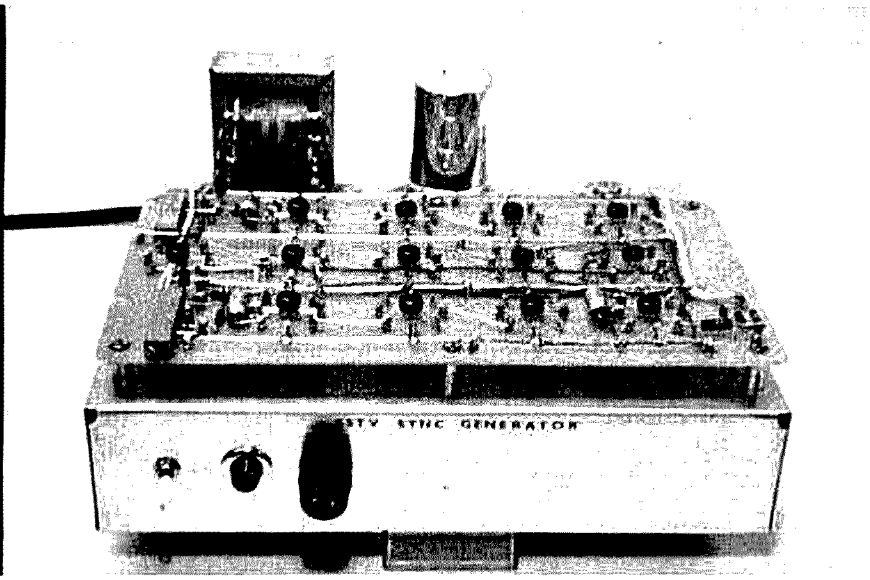

sync generator

for sstv

RTL flip-flops and gates in a circuit designed to provide stable horizontal and vertical sync pulses for slow-scan television

David R. Patterson, WA2EWO, 524 Gropp Avenue, Trenton, New Jersey

The availability of low-cost resistor-transistor logic (RTL) ICs suggests the possibility of constructing a stable sync generator for slow-scan television application.

Present sstv standards use a 15-Hz line-scanning rate and an 8-second frame rate. This combination gives the 120-line raster suggested under present standards. Normally these scanning rates are referenced to the 60-Hz power line frequency to provide stability and reduce 60-Hz hum in the sstv picture.

counting circuit

The counting-type sync generator shown here was built using two popular IC twins – the RTL 914 dual 2-input positive-logic NOR gate and the RTL 923 J-K flip-flop.* The J-K flip-flop is a very convenient device for use in counting circuits. One input pulse changes the output state, while a second pulse is

*Don't overlook the HEP line of ICs for this and similar projects. The HEP series are selected "universal" substitutes chosen from Motorola's extensive family of devices for "Hobbyists, Experimenters, and Professionals" – hence "HEP". For example the 914 gate (Motorola MC714G) is available as HEP584. It will perform as indicated in this application. editor.

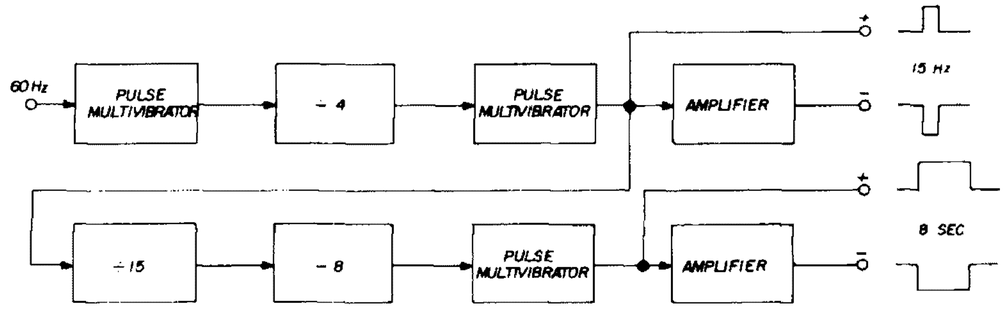

fig. 1. Block diagram showing counting sequences to obtain sstv horizontal and vertical scanning pulses.

necessary to restore the output to its original state. Thus two input pulses are required to complete one output cycle of the flip-flop. The counting process may be expanded by adding more flip-flops in series.

The output frequency of the flip-flop chain is related to its input in terms of powers of 2. This action is expressed as 2^n where n is the number of flip-flops connected in series. Flip-flops interconnected in this way form what is known as a binary ripple counter. If it's desired to count by a sequence other than 2^n, the normal counting sequence of the binary counter must be altered to achieve the desired count. Such change may be obtained by several methods. In the sync generator described here, gates are used to sense a particular count. The gates are then reset to restart the first counter so that the sequence begins again.

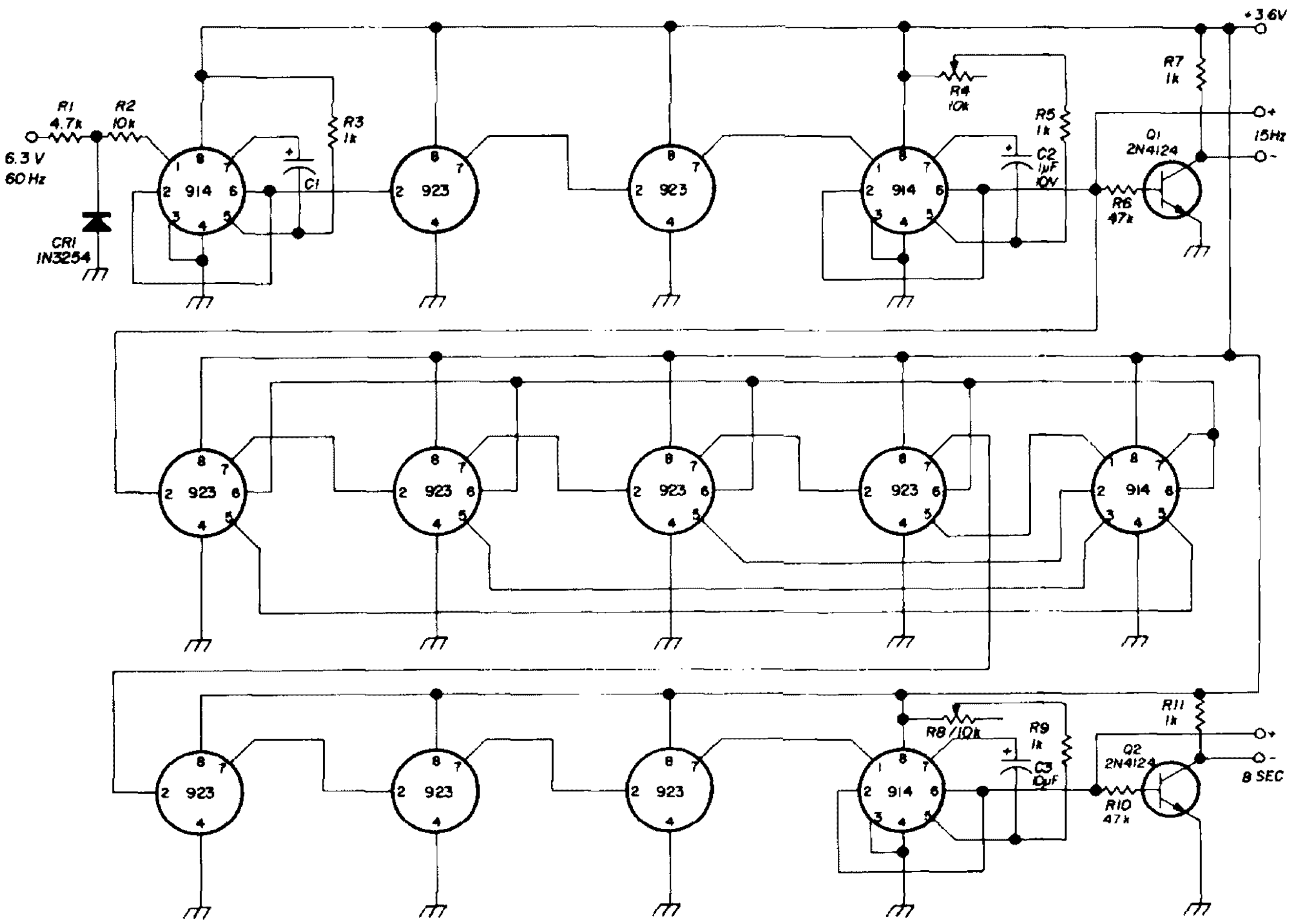

fig. 2. Sync-generator schematic. Binary ripple counters and 2-input NOR gates provide desired output.

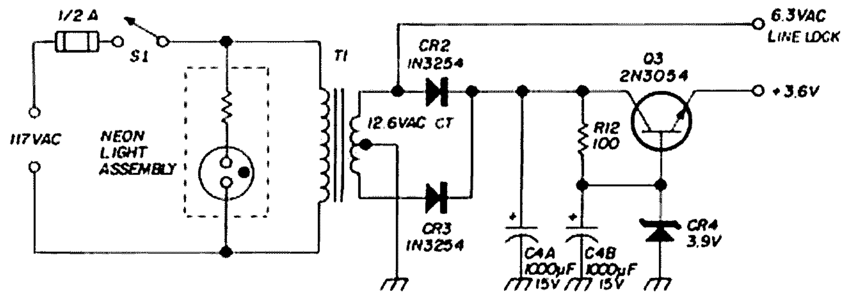

fig. 3. Suggested power supply. Regulation is provided by the zener in the base of Q3.

counting sequence

The block diagram of fig. 1 shows three basic counting sequences in the sync generator. The reference frequency, 60 Hz, is divided or counted by 4. As explained previously, this process will require two flip-flops; i. e., $2^2 = 4$. This circuit provides one output pulse for four input pulses, or an output frequency of 15 Hz. The next counter is a basic "divide by 16" counter that has been modified by gates to count by 15. Two gates (one IC package) are parallel-connected to sense the output of each J-K flip-flop and reset on the 15th input pulse. Thus the output is 1 Hz. The final counting sequence is provided by a "divide by 8" counter. Since $2^3 = 8$, three flip-flops and no gates are required. For every 8 pulses one output pulse results, or one output pulse every 8 seconds. This provides the sstv frame rate described above.

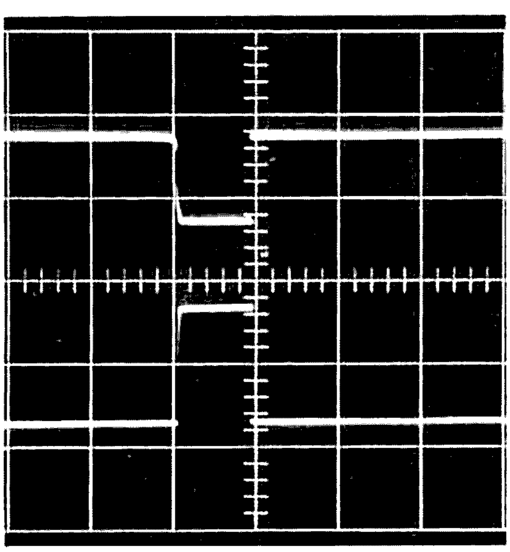

fig. 4. Oscillograph of the 15 Hz scanning pulse.

To provide the desired output-pulse width, two gates (one IC package) may be connected as a monostable multivibrator. The multivibrator output pulse can be adjusted to the desired pulse width by changing the RC time constant and bias voltage, or both, applied to the gate. Two units supply the proper output-pulse width of 5 ms for the 15-Hz scanning rate and 30 ms for the 8-second vertical or frame rate. A third multivibrator supplies a fast-fall-time pulse, derived from the power-line frequency, to trigger the first counter. The complete sync generator is shown in fig. 2.

power supply

Both output-pulse polarities are indicated on the schematic. Individual preferences will determine the actual polarity used. A simple emitter follower, zener-referenced power supply is shown in fig. 3.

adjustments

Only two adjustments are required. The output-pulse width is adjusted by R4 to provide the 5-ms output pulse at a 15 Hz rate, and R8 is adjusted to provide a 30-ms pulse every 8 seconds. Fig. 4 shows the plus and minus output of the 15-Hz, 5-ms line-scanning pulse. Positive output is 1 volt, and negative output is 3.6 volts.

The circuit described has proved to be an economical way to obtain stable synchronizing pulses for sstv use. The breadboard-type construction allows easy access to any point in the counting circuits if pulses other than those for sstv are desired.

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getting started in microwaves

Buy new,
build your own,
or scrounge surplus;
here is how
to start experimenting
with microwaves

William T. Roubal, 2725 Montlake Boulevard East, Seattle, Washington 98102

Many potential microwave enthusiasts have probably been frightened away simply because of the high price of new equipment. Such fears need not prevail, however, since many quality waveguide components, tubes and auxiliary pieces are available from dealers in surplus, often for as little as 1% of the new-market cost. Additionally, if you are mechanically inclined and can do metal work, you can fabricate some really beautiful pieces of your own equipment. Homemade components may not be devices calibrated to the nth degree, but they will be quite functional and will illustrate basic principles.

This article, therefore, is to show that you don't need unlimited resources to do some very interesting things with microwaves. Accordingly, nothing in the way of theory will be presented here. There are numerous applications for microwaves, both in communication and scientific research. The beginner as well as the advanced reader will find the public and college libraries the best source of information on both theory and applications.

getting started

Pick up an inexpensive, low-power

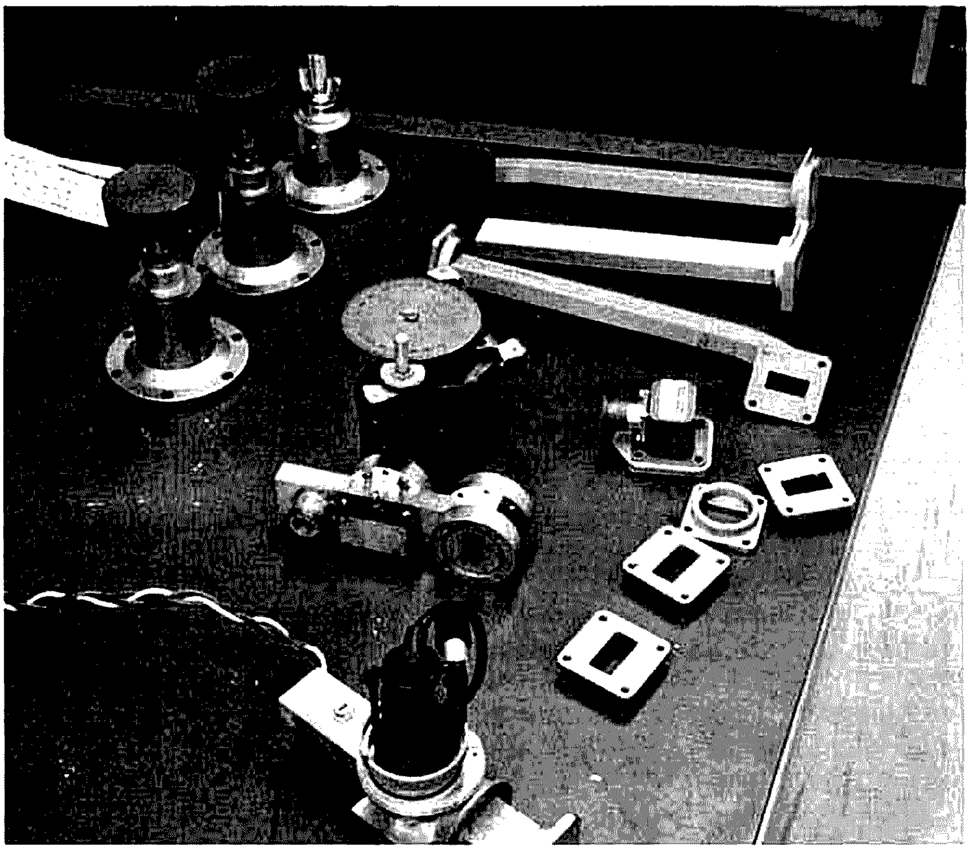

fig. 1. X-band waveguide components. From bottom to top are surplus 723 A/B klystron and holder, surplus resonant cavity, straight and bent sections of waveguides, UG-39/U waveguide flanges for do-it-yourself construction projects and waveguide-to-coax adaptor (next to flanges). The peculiar looking objects at the upper left are waveguide supports and were made by turning down the bakelite knobs salvaged from broken pressure regulators. The total value of the components if purchased outright would exceed $1,000. From surplus, all were obtained for less than $50.

klystron tube, hook it up to an appropriate power supply, and you're generating microwaves – no need to worry about interelectrode capacitance limiting your frequency, adjusting external tank circuits or watching lead dress, for all of these, and more, go out the window when you use a klystron. Add a few more inexpensive surplus items (or use homemade devices) and you're off and running. You can gradually expand your projects as your knowledge and technique develop.

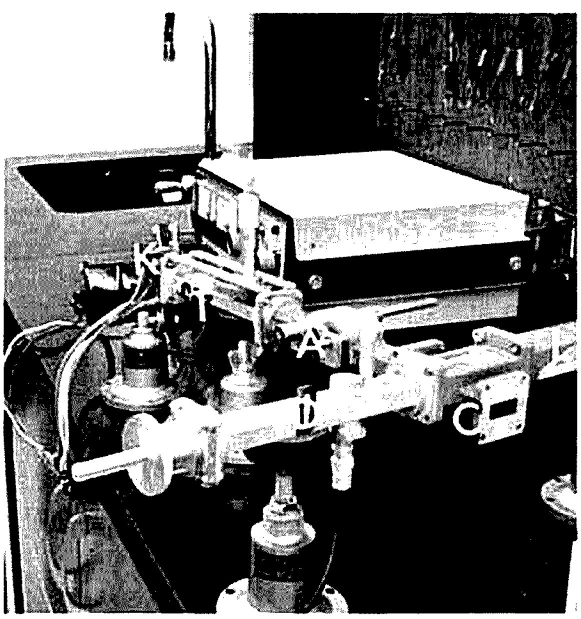

fig. 2. Experimental X-band microwave string.

Waveguides and related "plumbing" found most commonly are for X-band (9 GHz) applications (fig. 1). Fortunately some of the least expensive klystrons are also X-band generators. Accordingly, I shall limit my discussion to X-band components and methodology.

Fig. 2 shows a microwave string consisting of a Varian 998044 klystron (K), cooled by a small squirrel-cage fan (always force cool even the smallest klystron), homemade slide screw tuner (SST), surplus attenuator (A), surplus directional coupler (C), surplus absorptive

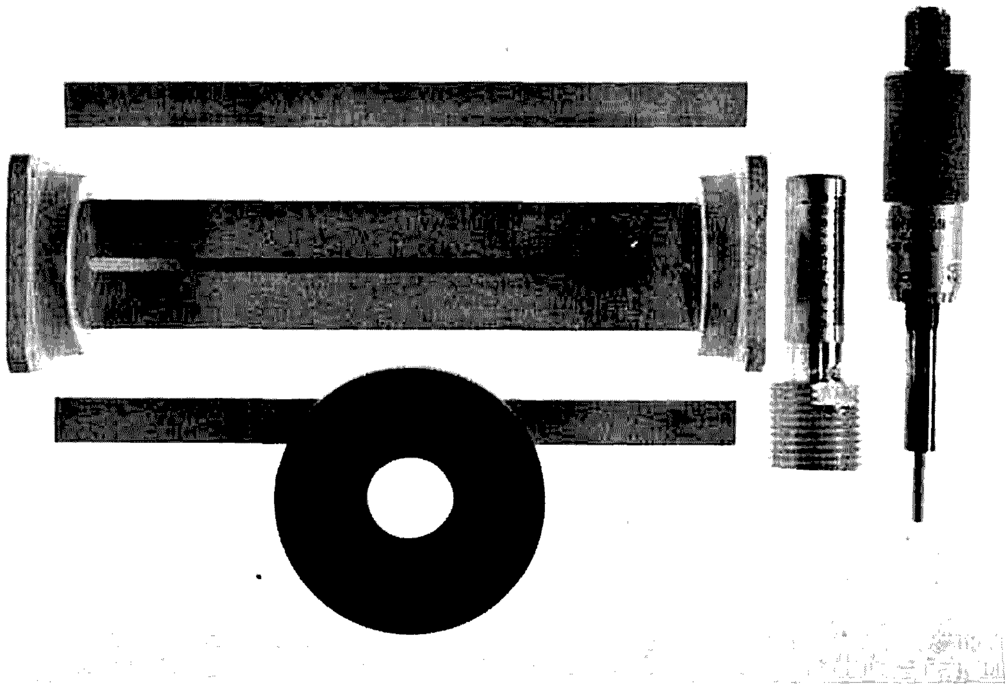

fig. 4. The homemade slide screw tuner. The tuning mechanism is a Sears micrometer body. In this version the threaded cylinder which holds the micrometer is ¾-inch OD instead of the specified 1 inch. The 1-inch piece is better electrically as it presents a larger surface area upon which to slide. It is included in fig. 5.

load (L) and a homemade tunable crystal detector (D). What you connect beyond the directional coupler will depend on what you want to study or illustrate. The assembly of fig. 2 is used for studying

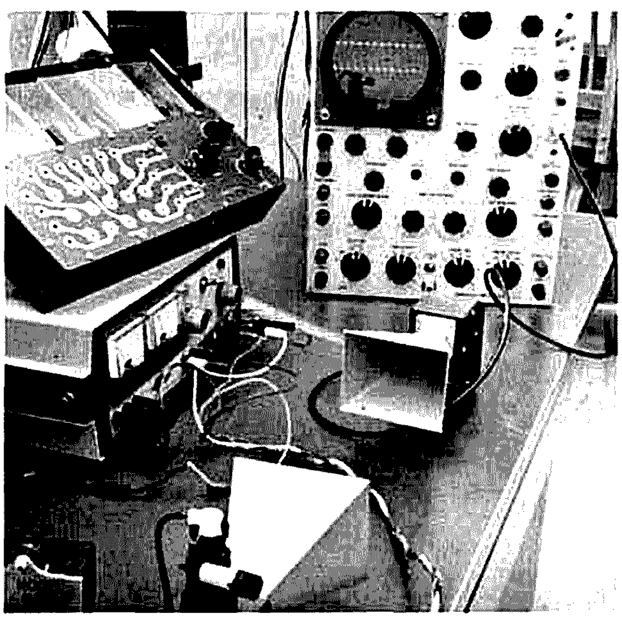

fig. 3. Microwave transmitter and receiver. The Berkeley Physics Lab by Heath-kit — transmitter and receiver (right foreground) and experimental patch board (sitting on power supply) — can be purchased prefabricated and complete, or ordered a part at a time, obtaining only what you need.

dielectric properties of materials which are inserted into a short section of empty waveguide and attached to the open port of the coupler. A simpler setup might consist of a klystron and horn radiator (transmitting antenna), and, at some distance away, a receiver (another horn coupled directly to a crystal detector). Such an arrangement can be used for demonstrating microwave propagation polarization and reflection. Fig. 3 shows one such setup using Heath equipment. Here a square wave is being used to modulate the klystron (achieved simply by coupling a square wave source to the repeller voltage source using a 0.01 μF coupling capacitor.)

equipment assembly

Since the exposed metal envelope of the reflex klystron tube forms the anode, the tube is normally operated with the anode grounded and with the cathode and repeller operated below ground potential. For either the Varian klystron shown in fig. 2 ($300, new) or the more common 723A/B (or 2K25) shown in fig. 1 (about $30, new), you will need a

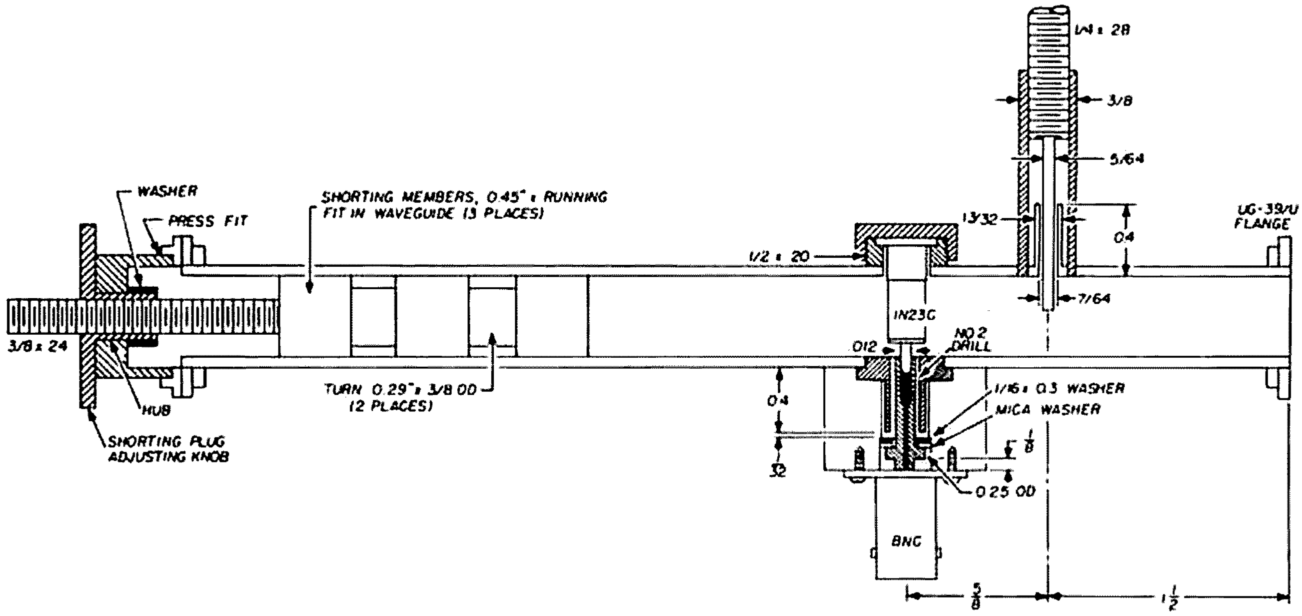

fig. 6. The homemade tunable crystal detector. All parts are brass plated with LT-26 electroless tin and soldered together.

well-regulated variable 0-300 volts dc cathode-to-anode supply, plus a variable -300 volts dc supply for the repeller; filaments, as usual, are 6.3 volts. The Heath Model IP-17 power supply provides all three voltages.

Though not an absolute necessity, the slide screw tuner will allow you to match up the microwave string for maximum overall efficiency. Used tuners, when available, may run $20 or more ($250, new), however, the metal working reader can build his own for about $11 (figs. 4 and 5). Variable attenuators may cost anywhere from $100 to $500, new. Nevertheless, the author has recently seen very good quality surplus units for $15. Similarly, brand new directional couplers will probably be priced under $20. Fig. 6 shows a cross-section of my homemade crystal detector, and fig. 7 is a photo of the parts assembly.

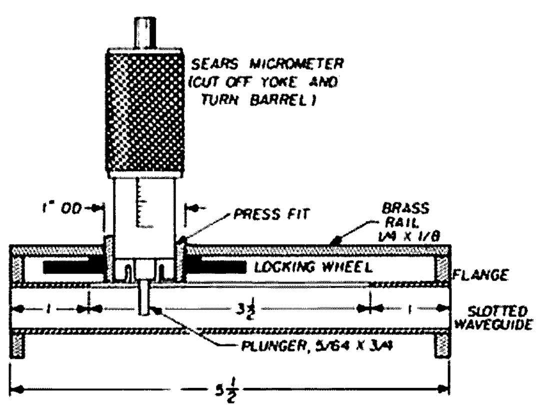

fig. 5. Construction details for the slide screw tuner. All fabricated parts are brass plated with electroless tin, using a simple, one-step process, before soldering together. Note the slot is shown in fig. 4 extending almost to the ends of the guide have been modified in this drawing. This is the recommended construction.

surplus

If you're lucky, your surplus items will be either gold or silver plated. If gold plated, they may show little or no tarnish. Usually a cleanup with nonflammable dry cleaning solvent followed with a soapy water scrubbing of exterior surfaces will make metal parts shine like new. Silver plated components, on the other hand, may be so badly tarnished as to look almost black. Carefully disassemble as much of the unit as you can (use caution since some items such as

*Both electroless tin and electroless gold are available from Shipley Company, Inc., 2300 Washington Street, Newton, Massachusetts 02162. More information on electroless plating is available in the article by Larry Hutchinson, "Practical Photofabrication of Printed-Circuit Boards," *ham radio*, September 1971, page 17.

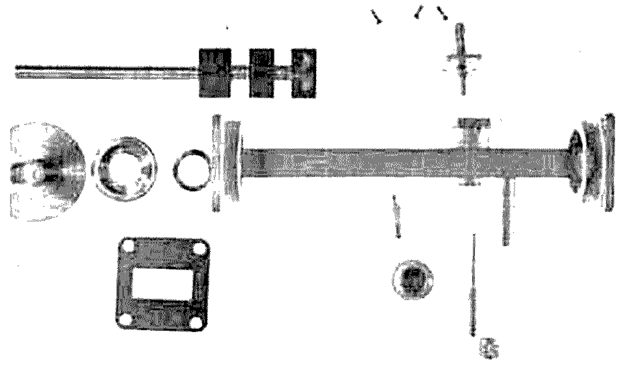

fig. 7. Construction details of the tunable crystal detector. The movable short and the tuning screw are adjusted for maximum crystal output. The drawing of fig. 6 differs in a few dimensions from the unit shown here. The text includes a few construction suggestions and hints on plating the parts. The real perfectionist can plate all the parts with gold.

slotted lines and certain couplers are very carefully mated together during assembly and are best left intact) and remove the tarnish using any of the ordinary liquid dip tableware tarnish removers.

Many used components will be made entirely of copper, bronze or brass without plating; more often than not, tarnish, corrosion and old paint are to be expected. Heavy tarnish, pitting and old paint can be removed from exterior surfaces with fine grit, wet-dry sandpaper and water. Generally, the internal surfaces will show less imperfections than the outside. Unless absolutely necessary, do not apply sandpaper or files to internal surfaces because close tolerances should not be altered. After the piece has been cleaned up, etch the piece briefly in mild acetic acid (or very dilute nitric acid). Rinse and dry the part, and plate the entire part with electroless tin.* The tin plate will improve electrical conductivity and will add new life to the metal. Tin-plated metal to be soldered can be heated in a direct flame; the plating will prevent oxidation and discoloration. For the ultimate in conductivity and corrosion resistance, brass parts can be gold plated with Shipley's electroless gold.

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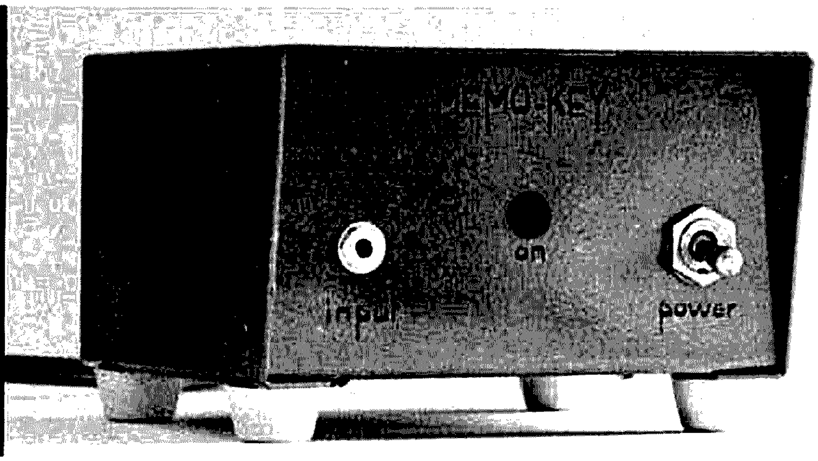


memo-key

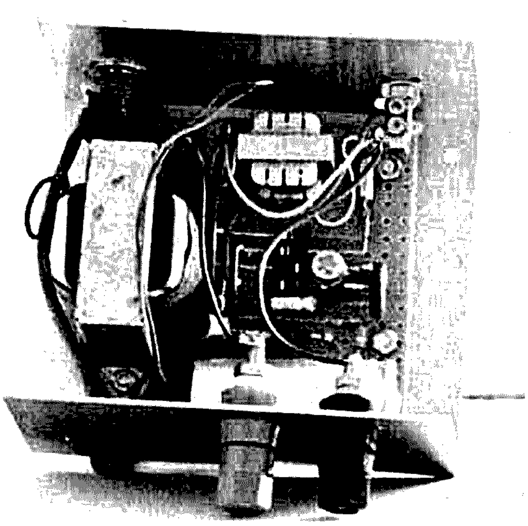
An inexpensive tape recorder and this electronic switch will increase your enjoyment of CW and RTTY

James R. Huffman, WA7SCB, Box 307, Mountain View, Wyoming 82939

This article may revolutionize your thinking about using code wheels for CW operation and paper tapes for RTTY. The memo-key can enhance your operating pleasure in many ways. Applications of the design are numerous and are limited only by your imagination.

how it works

The memo-key is shown in fig. 1. The circuit detects the presence of tones of various lengths that have been prerecorded on tape. These tones, which may be the recorded output of a code oscillator, side-tone from a receiver or keyer, or even a whistle, are converted to



Suggested parts layout.

on-off switching impulses in the output stage. The device will switch a maximum of 100 mA at 40 volts, which is sufficient to actuate a CW keying relay or drive a grid-block-keyed transmitter directly. Other applications are described later.

The memo-key may be operated from batteries or a power supply, as shown in fig. 1. Current drain is low, because the high-current output switch is not connected across the power supply. When driving a relay, a diode should be placed across the relay coil to suppress high-voltage spikes. The value of capacitor C4 may have to be changed slightly when switching currents other than 100 mA. The discharge time constant of the circuit is set for switching 100 mA; any load less than 100 mA slows the switching action considerably. In fact, when switching a low-current RTTY load, you may wish to eliminate C4 entirely, although sensitivity will decrease.

Two modifications may be necessary if you wish to use the memo-key with a tape deck. A level control should be connected across the deck output, then to the memo-key input. Also, no transformer will be needed at the input (T1 in fig. 1).

construction

Building the memo-key is no problem. A suggested layout is shown in the photo; however, layout isn't critical, nor is packaging. You may wish to use the utility box shown; rack-mount the unit, or even install it in the same enclosure with an RTTY converter. If you'd like to build a code oscillator, try one of the small 2-transistor units available from Burstein-Applebee or Allied Radio Shack. These units are small, operate from 6 Vdc, and are inexpensive.

As for the tape recorder, a good rule-of-thumb is: your existing tape recorder will probably be satisfactory for the applications shown. If you must purchase a tape recorder for this project, bear in mind the convenience of filing cassettes. Some of the catalog outlets send flyers showing cassette recorders for less than $20.00. Any of these recorders is satisfactory for use with the memo-key applications shown in this article.

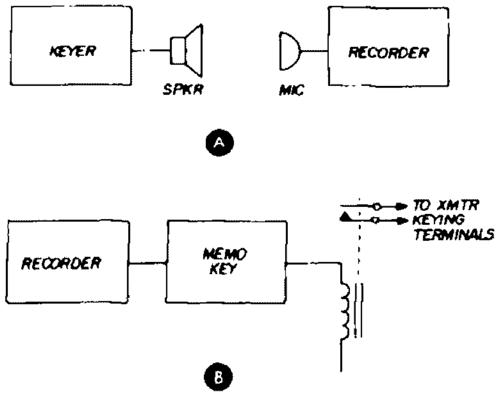


fig. 1. Memo-key schematic. Circuit is a simple electronic switch that will handle loads to 40 volts at 100 mA. Observe polarities when using suggested power supply or batteries.

some applications

Applying the memo-key to your particular needs is easy. For example, memo-key may be used as a code-wheel replac-

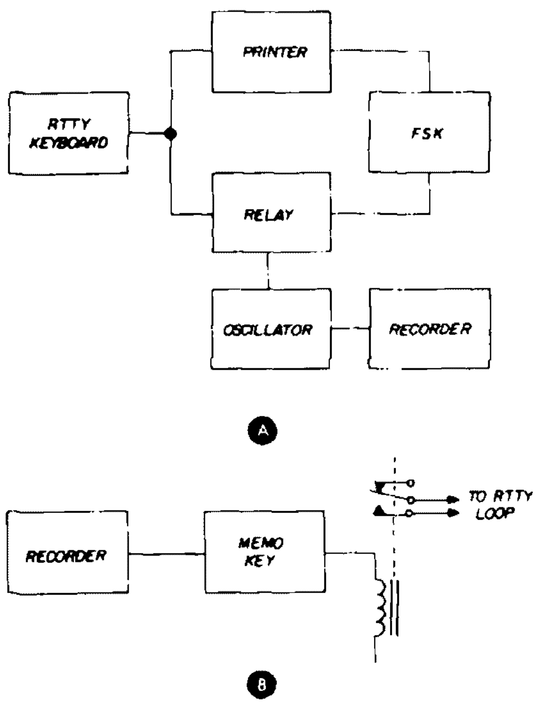


fig. 2. Using the memo-key for CW. Circuit keys the transmitter according to prerecorded input on magnetic tape. Reprogramming is easy, unlike paper tapes or "code wheels."

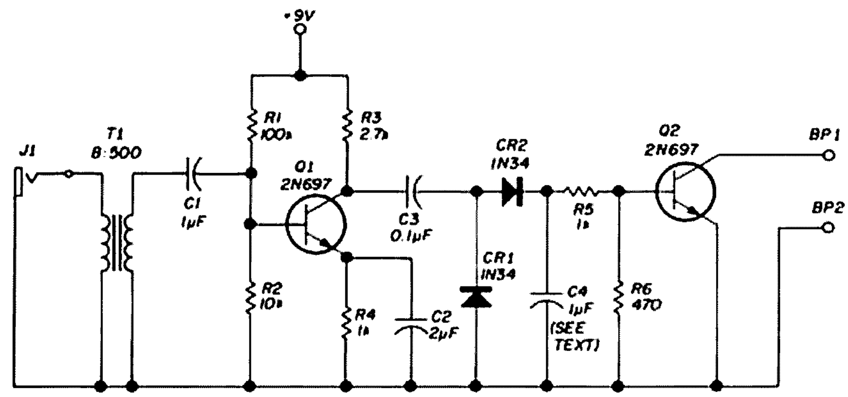
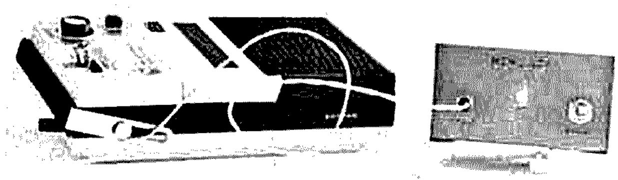


fig. 3. Memo-key for RTTY. Recorded input is stored on tape (mark = tone; space = no tone). Recorder output drives memo-key, which actuates a relay to drive the RTTY keyboard. To keep loop closed, the bias oscillator can feed a mark signal except in "play" mode.

ement to call CQ in contests. It may be used to close and open your end of a contact. In RTTY, you may use the memo-key to send "fox" tests, standard CQ's, elaborate ID's, or anything that can be put on paper tape. Other applications are also possible. Voiced "dits" and "dahs" can be converted to code-oscillator output for code-practice sessions.



Complete memo-key with cassette recorder. System is compact and offers easy storage of data for the active operator.

Control of transmitters, receivers, or other equipment in the home or amateur station is also possible—all information is remembered and stored as in a complex digital computer. You could push a "play" button on the tape recorder, and instantly the room lights would come on, then the transmitter and receiver tube filaments — all by feeding a shift register with the memo-key output.

concluding remarks

The advantages of using a system like memo-key are numerous: for one, programming is easily changed. The old code wheel is fine for calling CQ, but it doesn't lend itself to reprogramming. With memo-key, you can record yourself calling a station; then at the beginning and end of each transmission merely energize the recorder, and memo-key takes care of the identification.

Another advantage of memo-key is that tape storage is easily accomplished—a boon to the active RTTY and contest operator. A small cassette properly labeled would reduce storage space considerably. Since those paper tapes represent a lot of work it's to your advantage to use magnetic-tape storage, because magnetic tape is nearly impossible to rip, tear, or otherwise mutilate if the tape is neatly tucked inside a cassette case.

Wherever you put it; however you package it; however you use it—the memo-key will serve you well.

ham radio

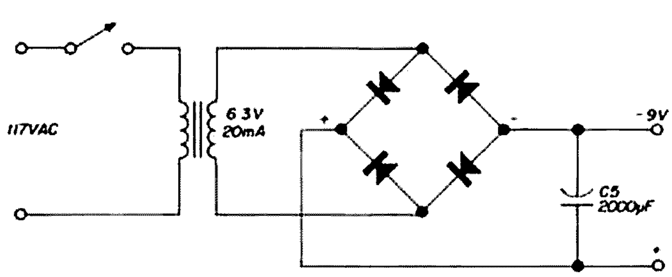
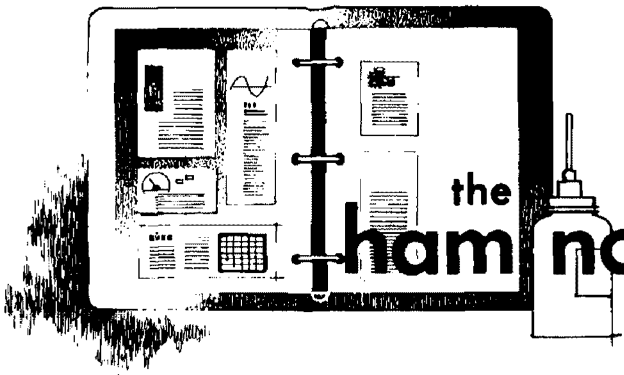


fig. 4. Suggested power supply circuit.



the ham notebook

horizontal or vertical?

Antennas are considered to be reciprocal devices, transmitting and receiving equally well. Common amateur practice is to use the same antenna for both functions. Under some conditions, though, better results can be obtained with separate antennas.

On 1.8 and 3.5 MHz, and to a lesser extent on 7 MHz, high-gain directional antennas are impractical for most hams. The average amateur low-frequency antenna is fairly omnidirectional. Therefore, the antenna can generally provide no improvement in the signal-to-noise ratio, when the principal noise source is atmospheric.

Long-distance communication requires low-angle radiation. A study of theoretical antenna patterns shows that a horizontal dipole has the best low-angle radiation characteristics when it is one-half wavelength above ground. Since this is 140 feet at 3.5 MHz, most amateur 3.5-MHz horizontal antennas radiate at predominantly high angles – fine for short skip, but not very good for the really long-haul stuff.

The radiation from a vertical antenna is mostly low-angle at any height less than 5/8 wavelength. Many hams, hoping to improve their low-frequency DX performance, have installed vertical antennas and radial ground systems. They have generally been disappointed; they did not hear any more DX than they did on their old horizontal antennas, nor was the DX they did hear any louder. Sometimes the vertical seemed to be worse.

There is a reason for this. On the low-frequency bands the chief limiting factor is noise. At any time, atmospheric and man-made noise have a local field strength with which the signal must compete to produce a usable signal-to-noise ratio at the output of the receiver. If the signal field strength does not have a favorable relationship to the noise field strength, there is nothing to be done. The improvement has to be made at the transmitting end – either by the distant station raising power, or by increasing the efficiency of *his* antenna. It might almost be said that the configuration of the *receiving* antenna on the low-frequency bands is unimportant. As long as the receiving antenna is long enough to produce a fairly loud noise output from the receiver, it might as well be a short, random-length piece of wire!

The transmitting antenna may actually be worse for receiving. This is likely to be true if the antenna is vertical. Vertical antennas are much more sensitive to man-made noise, which is predominantly vertically polarized. Even after point sources of man-made noise are eliminated, almost every amateur lives in an environment with a background of man-made noise that is difficult to distinguish from true static. The situation is worst on the low-frequency bands.

I have been reading about the sensitivity of vertical antennas to man-made noise for years, but it was not until I had both vertical and horizontal 7-MHz antennas available at the flick of a switch that I realized how great the difference was. Switching from horizontal to vertical usually resulted in a noise increase of about 10 dB!

Transmitting, it was a different story.

For example, from my location ZS1A is pretty good DX on 7 MHz. I was able to work him with relative ease using my roof-mounted trap vertical. Using my horizontal dipole, I was usually able to hear him with a better signal-to-noise ratio. But I was *never* able to work him using the horizontal for transmitting.

Later experiments working the Orient and the Pacific confirmed this. The horizontal was better for receiving, but the vertical was from one to three S-points better on transmit. The difference, of course, was the greater man-made noise pickup of the vertical. I did not have a vertical antenna for 3.5 or 1.8 MHz, but I am sure the results would have been comparable.

From my own experience and a study of antenna and noise theory, I believe the following principles can be established:

1. For extreme long-distance transmission on low-frequency bands, verticals should be better than horizontals and probably are.

2. Comparisons of the effectiveness of low-frequency verticals with horizontals on the basis of receiving tests are not always valid. Field-strength measurements should be made.

3. For low-frequency DX reception, the antenna is of little importance, but a vertical may be worse than a horizontal because of man-made noise.

Undoubtedly the best scheme is to have both vertical and horizontal antennas available, with switching to make it possible for either to be used for receiving or transmitting independently of the other.

Of course, these findings only apply when it is not possible to erect an antenna having worthwhile gain and directivity – the usual situation on the low-frequency bands.

Harry R. Hyder, W7IV

Swan 350 CW monitor

This simple circuit provides a built-in CW monitor for the Swan 350. It can be built compactly, installed quickly and has plenty of output.

The parts were mounted on a small printed-circuit board with the resistors installed vertically. I used a solder lug as a mounting foot, secured by one of the crystal filter mounting nuts. This location permits relatively short leads to the required locations.

The resistor, R6, provides isolation of the circuit to prevent loading of the 6GK6 grid circuit and to attenuate the oscillator output to a comfortable level. A value between 1 and 2 megohms should

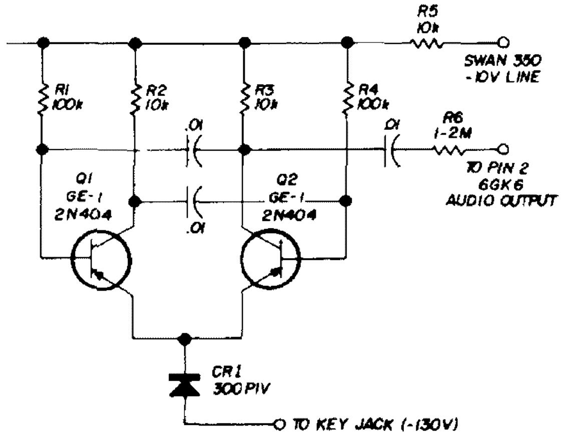


fig. 1. Simple CW monitor for the Swan 350.

do. With the components shown, the frequency is approximately 800 Hz. This frequency may be changed by altering any of the constants shown. The simplest way would be substituting various transistors. Just about all transistors will oscillate, and combining a germanium audio type and a pnp silicon rf type can provide some interesting notes!

With the transceiver in the receive mode and the audio and rf gain controls down, the oscillator can be used for practicing the code or adjusting a keyer. I use a pair of stereo headphones with the sections connected in parallel, and the note has more than enough volume. If you use a speaker, decreasing R6 to 1 megohm or 0.5 megohm should blast your eardrums. Changing the volume control during transmit will vary the output but will not change the pitch of the note.

Paul K. Pagel, K1KXA

new products

hf receivers

Radio Shack has introduced two new all solid-state 11-band receivers, the AX-190 for amateur band coverage and the SX-190 for general communications and shortwave listening.

Both receivers are dual conversion with a crystal-controlled first oscillator and tunable second oscillator. Sensitivity is given as 0.5 μV on ssb and CW, and 1.0 μV on a-m for 10 dB S+N/N.

A built-in Q-multiplier provides better than 60-dB image rejection and 50-dB spurious rejection. Built-in 25- and 100-kHz crystal calibrators assure visual dial accuracy of within 200 Hz. The dial reads direct to 1 kHz. Frequency stability is said to be better than 500 Hz per hour.

Other features include ceramic filters for sharp selectivity, dual time-constant agc, anl, crystal-controlled bfo, illuminated S-meter and a line/tape output and headphone jack. Each receiver has dual-regulated power supplies for operation on 120 Vac and 12 Vdc.

The AX-190 is crystal-controlled on 3.5-4, 7-7.5, 14-14.5, 15-15.5, 21-21.5, 27-27.5, 28-28.5, 28.5-29, 29-29.5 and 29.5-30 MHz, with a blank channel for special monitoring between 3.5 and 10 MHz. It has a vfo output and simplified interconnections for easy transmitter hook-up.

The SX-190 is supplied with crystals for reception on 3.5-4, 5.7-6.2, 7-7.5, 9.5-10, 11.5-12, 14-14.5, 15-15.5, 17.5-18 and 27-27.5 MHz, with blank channels for special monitoring on 3.5-10 MHz and 10-30 MHz.

The AX-190 and SX-190 receivers are priced at $249.95 each, with metal cabinet, 7 x 15 x 10-inches. A matching speaker is available for $19.95. The receivers are available at any of the local Allied Radio Shack stores. More information is available by using *check-off* on page 94.

regulated power supplies

The benefits of regulated power supplies are explained by Irving Gottlieb, W6HDM, in his new book "Regulated Power Supplies" published by Howard W. Sams. The first chapter lists the reasons why regulated power supplies are used. These include such benefits as increased efficiency, precision, greater dynamic range, feedback stabilization, higher signal-to-noise ratio and wider frequency response.

The second chapter discusses the static characteristics of regulated power sup-

More Details? CHECK—OFF Page 94

plies such as dc regulation, temperature effects, stability, long-time and short-time deviations and drift. Other chapters cover dissipation control and the uses of integrated circuits and monolithic modules. Representative examples of commercial power supplies are also explained. 160 pages; $5.95, softbound, from Comtec Books, Greenville, New Hampshire 03048.

fm transceiver with tone encoder

SAROC featured the introduction of Ross and White's new two-meter fm transceiver with built-in tone-burst encoder. The new RW-Bnd transceiver offers twelve-channel operation (four crystals supplied), three power levels of 0.1, 1 and 10 watts, all solid-state design, and the three-tone encoder. The tones are factory set for the three most common tone-burst frequencies, but they are easily changed in the field.

The unit sells for $359.95 with the built-in encoder. The same rig, but without the encoder, sells for $319.95. This model has provision for the installation of the encoder at any time in the future. Both units come with a microphone and mobile mounting bracket.

For complete specifications on this new rig, write to Ross and White Company, 50 West Dundee Rd., Wheeling, Illinois, 60090 and request the information sheet on the RW-Bnd. You can get the same information by using *check-off* on page 94.

hobbyist LEDs

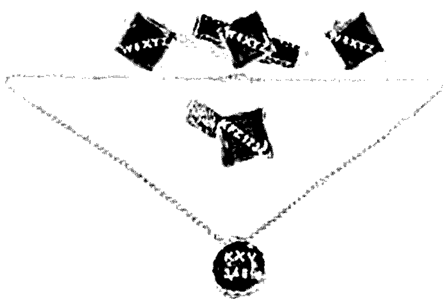
Priced for the amateur experimenter, premium-quality light-emitting diodes are now available through the national network of Sprague distributors. Each distributor will have a working display of many LEDs and a specially-selected inventory of LEDs and related gallium arsenide semiconductors including visible light sources, infra-red sources, indicator lamps, alpha-numeric readouts, photosensors and opto-isolators made by Monsanto.

Although intended primarily for the experimenter, the new line of LEDs are exact replacements for thousands of devices in standard equipment and for many conventional miniature lamps.

Sprague suggests numerous experimenter projects using the new devices stressing the low-current demands and exceptional long-life of the LEDs. Panel indicators, light communications and digital readouts are a few of the more conventional suggestions.

More information is available from Sprague Products Company, North Adams, Massachusetts 01247, or by using *check-off* on page 94.

jewelry



Hobby Jewelry offers a variety of earrings, tie clips, lockets and cufflinks custom made with any call sign, initial or photograph. A variety of sizes and a choice of gold or silver finish is offered, with both finishes guaranteed for a year. More information and prices are available from Hobby Jewelry, 13407 Shoup Avenue, Hawthorne, California 90250 or by using *check-off* on page 94.

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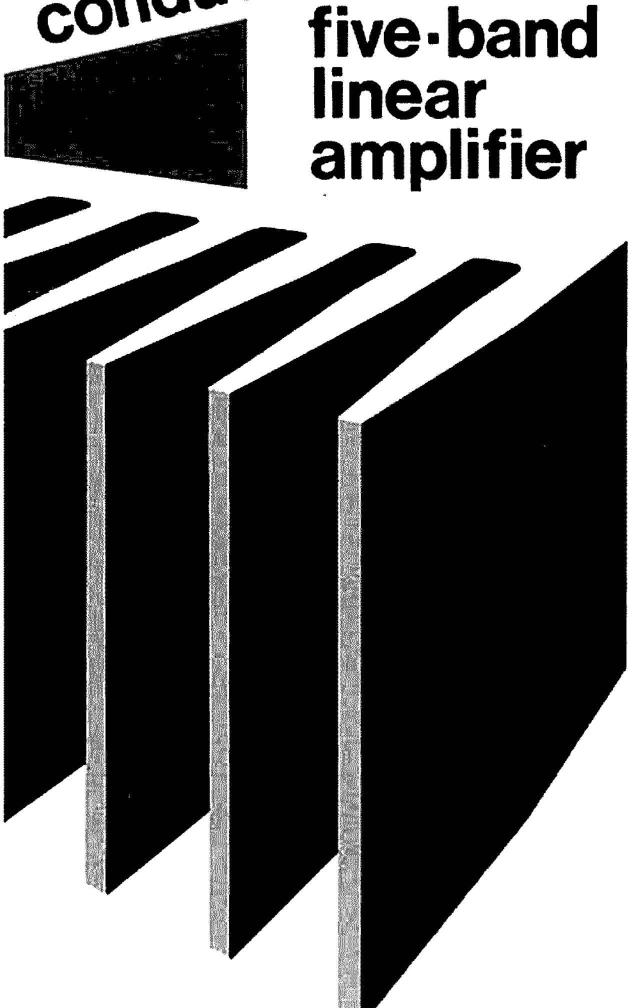
conduction-cooled five-band linear amplifier



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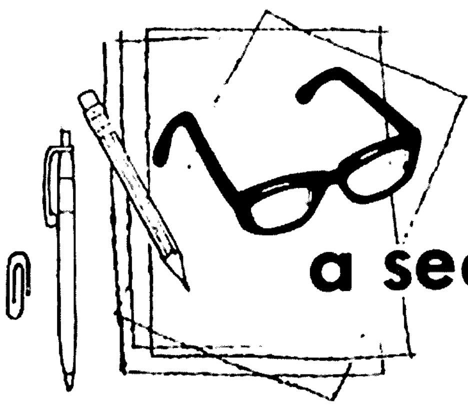


a second look

by jim fisk

Our society is a very mobile one, and when it comes to traveling long distances, most of us fly with one of the commercial airlines. It's only natural for the fm'er, with his portable fm rig, to question the possibility of using his equipment on commercial flights.

It is popularly believed that all we have to do is obtain the Captain's permission to operate; surely our little two-watt fm rig is not going to cause any interference with the high-powered radio equipment used on board the aircraft. However, this is not the case – according to the Federal Air Regulations, approval must be obtained from the air carrier (airline) and *not* the pilot in command. However, once approved by the air carrier, the permission of the Captain in command must also be obtained to operate equipment aboard a particular flight.

Shortly after World War II, portable Japanese fm broadcast receivers started appearing on the market, and passengers started using them aboard commercial flights. At the same time, aircraft navigation radios started doing funny things, and it didn't take long to determine that the interference was being caused by rf radiation from the portable fm receivers. The aircraft radios literally went wild, and at least two aircraft accidents have been attributed to interference of this type.

When it was determined that this interference was present, the FAA promulgated new regulations, paragraph 91.19 of the Federal Air Regulations. This paragraph states that *no* electronic device may be operated aboard a commercial airliner *except* heart pacemakers, voice recorders, hearing aids, electric shavers and electric watches, unless the device has been approved by the air carrier or operator. The regulation further states that *the captain of the aircraft does not have the authority to authorize such operation.*

Consider, for a moment, what might happen if such operation were allowed. Suppose you have been operating all across the country, and your plane is about to land. A passenger with a briefcase telephone sitting across from you has been watching you operate. About 10 minutes before landing, he decides to call his wife. Unfortunately, his telephone operates on a frequency right in the middle of the glide slope spectrum. As soon as his transmitter is keyed, the glide slope indicator cross pointer goes up or down, and the autopilot follows it. This could be disastrous.

As an airliner flies across the country, the pilot changes frequency every 5 minutes or so. If several fm operators are on the same flight, only one can talk at a time, so some may decide to switch to other frequencies. When you start to figure out all of the i-f and carrier frequencies of the aircraft radios, plus the

(continued on page 86)

five-band conduction-cooled linear amplifier

A high performance table-top linear that provides noiseless operation on all bands from 3.5 to 28 MHz

Richard I. Bain, W9KIT, 4915 Ridgedale Drive, Fort Wayne, Indiana 46815

The idea of adding a linear amplifier to your station becomes quite attractive after fighting the DX dog-piles on 20, foreign broadcast QRM on 40, or QRN on 80. When I started planning a high-power linear amplifier for my station, I wanted a table-top unit that was quiet, lightweight, stable and efficient, and that would dissipate very little power on standby. I also wanted an amplifier as uncomplicated as possible; the fewer parts there are, the fewer there are to fail.

The conduction-cooled linear shown in fig. 1 nicely fills all of these requirements. The conduction-cooled Eimac 8873 power tubes provide up to 1500 watts PEP input without a blower, so the amplifier is absolutely quiet.

The amplifier is housed in a table-top sized cabinet 24 inches long and 12 inches high. The unit dissipates little power on standby, and power output on 80, 40, 20 and 15 meters is 600 watts; on 10 meters power output is slightly less.

circuit

In the complete amplifier circuit shown in fig. 1, two vacuum relays are used to switch the linear in and out of the circuit. These two relays are controlled by a third, smaller relay which also shorts out the protective-bias resistor. The use of input and output switching allows use of the linear with transceivers. The relays can be energized only when the high-voltage power supply is turned on.

The parasitic choke is located in the

cathode input circuit, and seems to be just as effective as if it were placed in the plate circuit.[1] The cathode and filaments are isolated from rf ground by a 3-winding rf choke wound on a ½- by 5-inch ferrite rod. The winding, consisting of two no. 16 wires for the filaments and one no. 20 for the cathode, fills the full length of the ferrite rod.

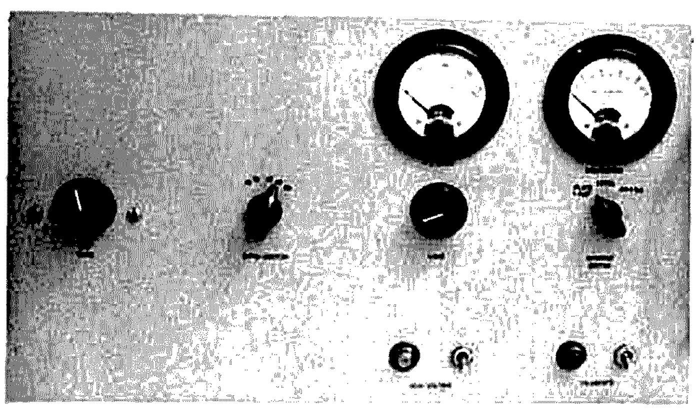
The 8.2-volt zener diode provides cathode bias to hold the two tubes to a total cathode current of approximately 50 mA. I used a 10-watt diode, which should be sufficient for up to 1500 watts input; for 2-kW operation, a 50-watt zener would be preferable. The resistor in parallel with the zener provides insurance against damage due to a floating ground in case the zener opens up. The 10k resistor in series with the zener diode provides cut-off bias during standby.

The rf plate choke I used is a surplus pie-wound unit. The National R-175A or B&W 800 are suitable substitutes. The plate switch has three sections, but only two are used. One switch section selects taps on the coil and adds extra loading capacitance on 3.5 MHz. The other section adds capacitance in the plate side in the 3.5-MHz position.

The plate-tuning capacitor is tapped down on the ten-meter coil. This reduces the effective minimum capacitance of the plate-tuning capacitor so more inductance is placed in the circuit when operating on the 28-MHz band.

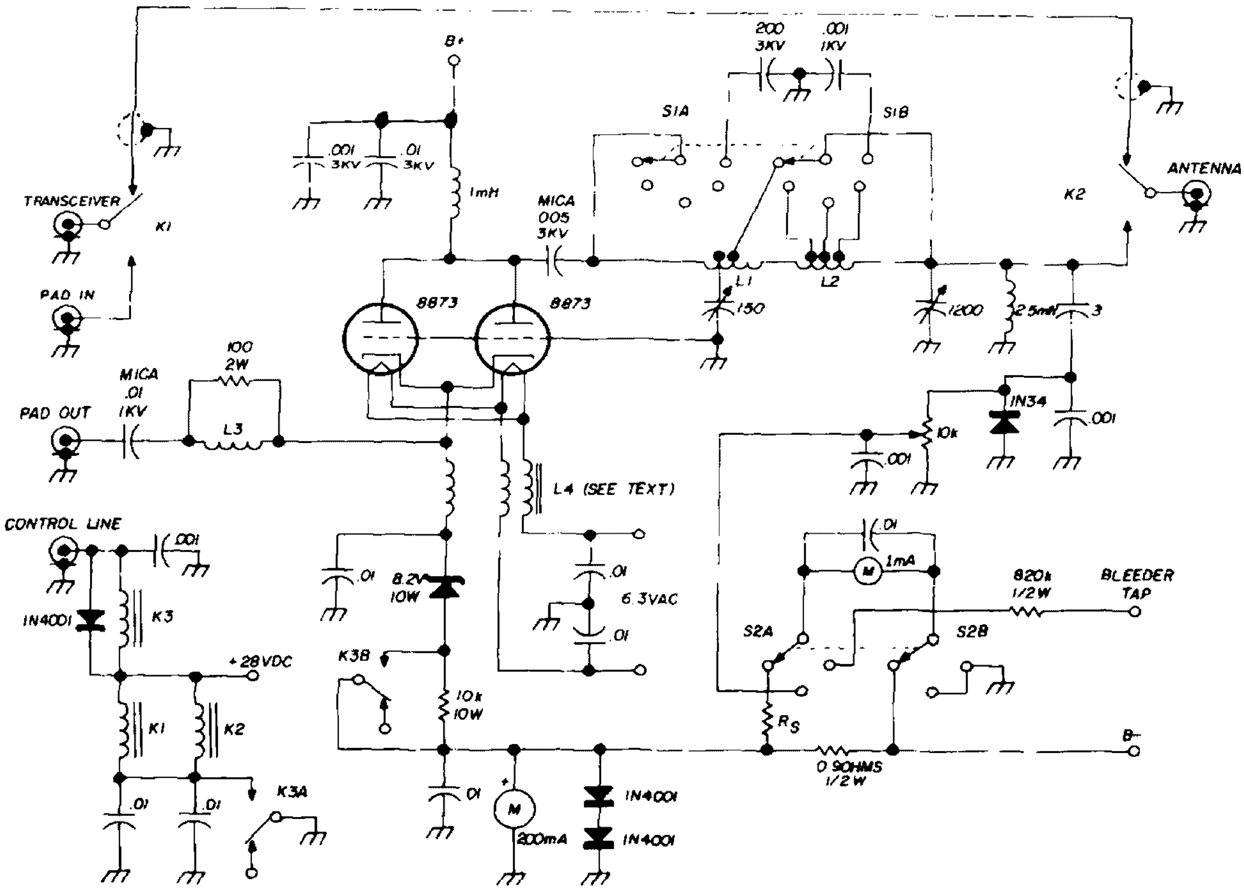
I wanted to run 1500 watts PEP on ssb, but this raised the question of obtaining reasonable tuned-circuit Q at both 1000 and 1500 watts input. However, I found that, by selecting a Q of 15 at 1000 watts, the Q was still 10 at 1500 watts; both fall within a satisfactory range. If you want to run 2000 watts PEP for ssb, the tank coil should have separate taps for CW and ssb.[1]

A second problem that had to be considered was the high fixed capacitance in the plate circuit due to tube output capacitance and stray capacitance to the heat sinks. Unfortunately, the total of this stray capacitance is sufficient to resonate the desired tank inductance on 10 meters. Therefore, I had to settle for a smaller coil, and consequently, higher Q. The sacrifice is small, however, and the 10-meter coil does not heat excessively during normal operation.



The simple front panel of the conduction-cooled linear. The front panel of the finished amplifier has handles, facilitating moving, handling and servicing the unit.

I debated about adding tuned circuits for cathode matching to gain the 1 or 2 dB advantage in intermodulation distortion, but decided to try the amplifier without them and see how it worked out. The cathode input impedance is in the 60- to 70-ohm region, a reasonable match for most exciters. A problem that popped up while using the linear was that exciter drive dropped off as a transmission progressed. I added a power-dropping attenuator pad so the exciter could be operated



L1	5 turns 3/16" copper tubing, 1¾" diameter, 2" long, tapped at 4 turns
L2	Air-Dux 2408D4 tapped at 1, 4, 7 and 13 turns from coarse-wound end
K1,K2	28 Vdc vacuum relays
K3	miniature 2-pole 28-Vdc relay

Heath sinks are 6" pieces of extruded stock, Thermal Associates no. S-511

fig. 1. Schematic diagram of the five-band 1500-watt conduction-cooled linear amplifier. Resistor Rs is chosen to give 1 ampere full-scale deflection.

at normal input levels where the output remains stable. The attenuator circuit is shown in fig. 2.

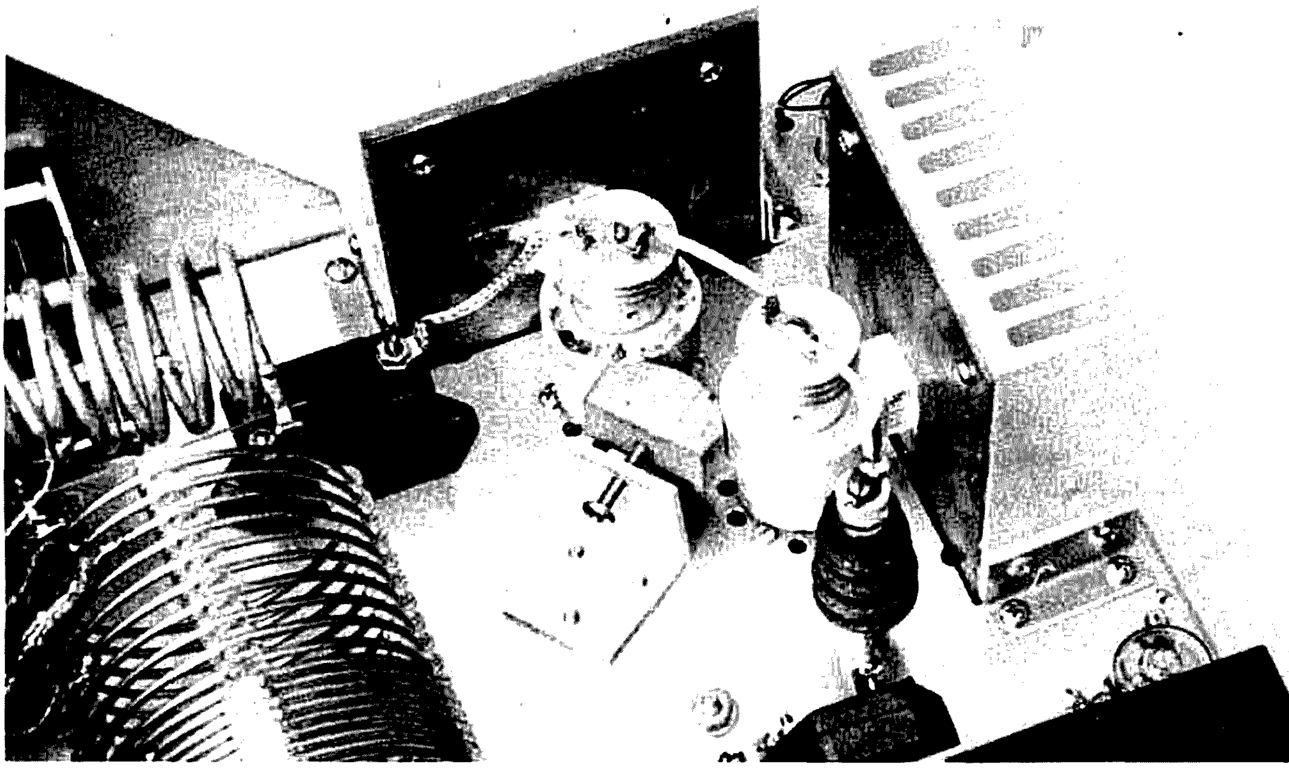
The attenuator is housed in a 2-by-3-by 5-inch minibox, with numerous ventilation holes. The pad could have been built on the amplifier chassis, but by being able to patch it in and out, it can be replaced by a coax jumper for operation with low-power exciters. The amplifier only requires about 16 watts drive for 1000 watts input on the lower bands, and slightly more on the higher bands.

One advantage of the pad is that the exciter sees a resistive 50-ohm load. Also, the pad tends to swamp out variations in cathode impedance due to plate tuning, so the exciter doesn't have to be returned each time the linear is adjusted.

construction

The amplifier shown in the photos was built on a 17- by 13- by 3-inch chassis with a panel 22 inches long by 11 inches high. The panel size was chosen to fit a surplus cabinet I had on hand. The chassis is supported by two side brackets to prevent excessive flexing.

The heat sinks used to cool the tubes are 6 by 4 inches with 1½-inch cooling fins. Thus, each heat sink provides 36 cubic inches of cooling volume. The heat sinks are faced with 1/16-inch copper to provide improved heat transfer from the vicinity of the thermal link blocks to the rest of the heat sink. The heat sinks are sub-mounted 1½ inch below the top of the chassis so the thermal links are



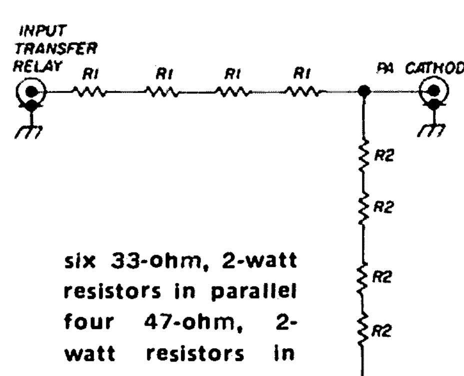
Close-up photograph shows the details of the final-tube mounting. Careful alignment of the tubes, heat sinks and clamping block is necessary to assure efficient and equal cooling of both tubes. To assure adequate ventilation of the heat sinks, drill plenty of holes in the equipment cabinet.

centered. Slots on the chassis sides beside the heat sinks allow free air movement up through the cooling fins.

The tube socket mounting holes are elongated to allow alignment of the flat face on the tubes with the heat sink surfaces. Also, the mounting holes in the aluminum angle brackets that hold the heat sinks in place are oversized – this allows some tilt adjustment of the heat sinks.

The alignment of the tubes and sinks must be nearly perfect, or the thermal link blocks will not sit flat on the sink, and heat transfer from the tubes to the heat sinks will be impaired. Thermal compound should be used liberally between the tubes and thermal links, between the thermal link and copper plate, and the copper facing.

The 8873s are held tightly against the heat sinks by a semicircular piece of diallyl phthalate, which is clamped in place by a number-10 screw through a hole tapped in a piece of aluminum angle bracket. The material used for the clamping block must be a good electrical insulator and be able to take temperatures up to 200° C. A tapered ceramic insulator might be used for this purpose. The chassis holes for the angle bracket are slotted so the bracket can be rotated to properly align the clamping block to

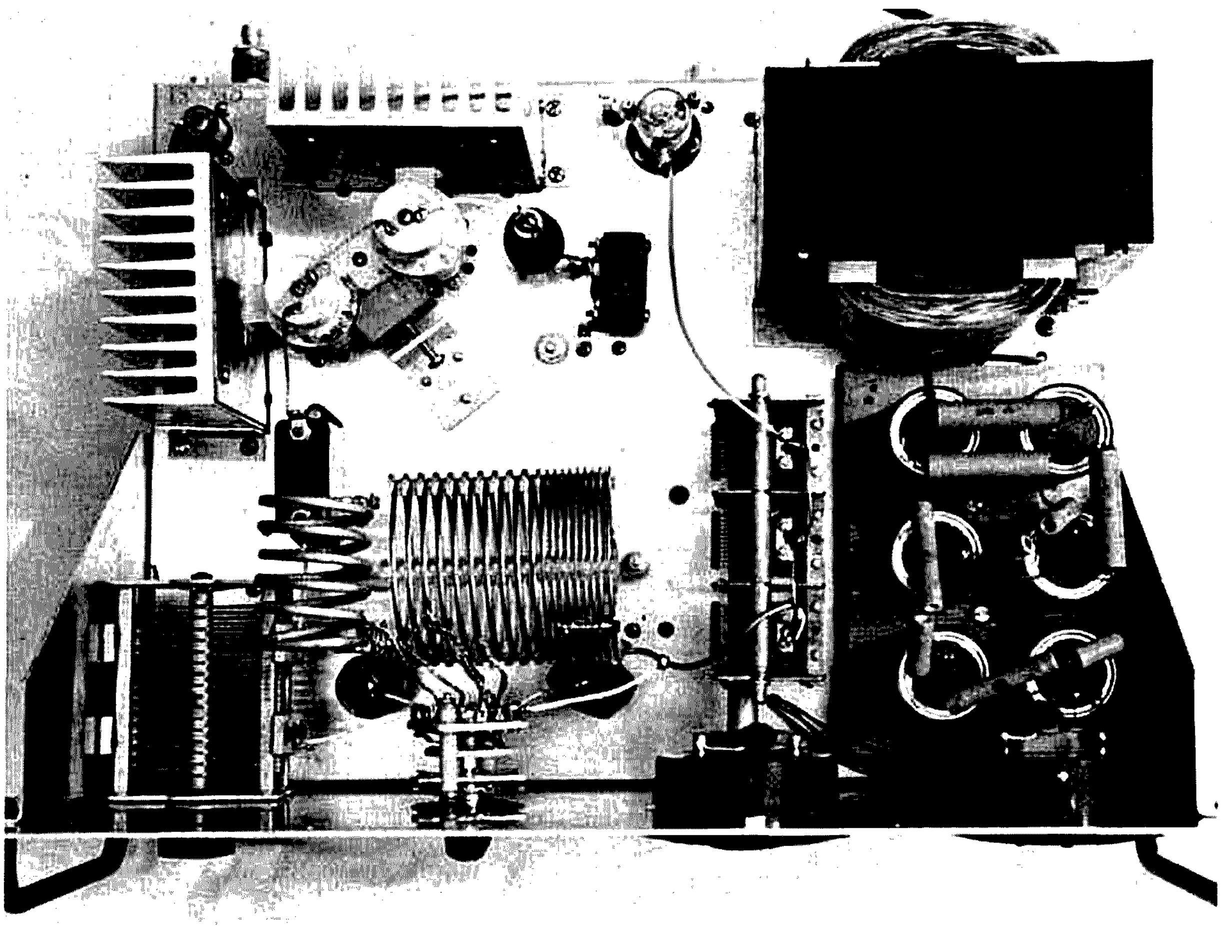


R1	six 33-ohm, 2-watt resistors in parallel
R2	four 47-ohm, 2-watt resistors in parallel

fig. 2. Power attenuator for use with high-power exciters. The amplifier requires only 16 watts drive for 1500 watts PEP input.

insure equal pressure on the two tubes.

It would, of course, be possible to mount the heat sinks in-line and clamp each tube individually with a toggle clamp.[2] However, regardless of the mounting you use, external air must be able to enter the cabinet, pass through the heat sinks and exhaust, all with ease. I put several rows of ¼ inch holes in the cabinet above and below each heat sink. The cabinet must be mounted on at least ¾ inch feet to allow air to enter through the bottom.



Overview of the complete amplifier and power supply show the clean layout and overall simplicity.

power supply

The high-voltage power supply has six series-connected capacitors for a total of 21 μF. This amount of filtering is adequate for up to 1500 watts PEP, but if you want to run 2000 watts input, more capacitance should be used. The rectifiers are 5000-volt bridges connected to give a 10 kV PIV diode array on each side of the transformer. The high voltage drops about 400 volts with a 500-mA load when operating from 117 Vac. Operation from 220 Vac should result in better regulation due to reduced primary current.

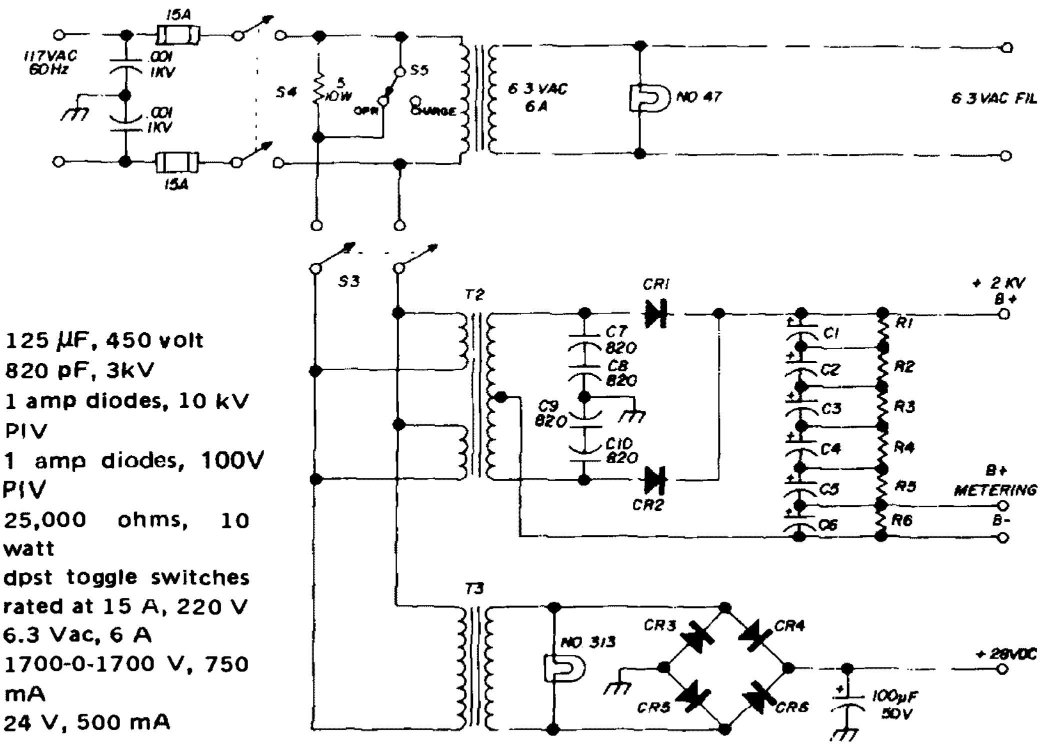
My diode rectifiers generate a small amount of hash, probably due to switching transients. Capacitors on the primary and secondary side of the power transformer keep most of the hash out of the 117-Vac line. Filter chokes may be needed for tougher cases.

The bypass capacitors also protect the diodes from large amplitude, narrow spikes which may be present on the ac line. The 28-volt supply provides power to the high-voltage indicator and the relays (see fig. 3).

The high-voltage supply is metered by tapping off the top of the bottom bleeder resistor. Since the bleeder resistors are not precision types, the actual value of

the high voltage should be measured by a more accurate means to obtain a correction factor; alternately, the value of the series-dropping resistor may be selected to make the meter read correctly.

The five-ohm resistor and S5 were added after six months of operation. I found that the surge current through S3 caused the switch contacts to stick closed. Closing S5 after S3 protects S3 and the high-voltage diodes.



C1-C6	125 μF, 450 volt
C7-C10	820 pF, 3kV
CR1, CR2	1 amp diodes, 10 kV PIV
CR3-CR6	1 amp diodes, 100V PIV
R1-R6	25,000 ohms, 10 watt
S2, S3	dpst toggle switches rated at 15 A, 220 V
T1	6.3 Vac, 6 A
T2	1700-0-1700 V, 750 mA
T3	24 V, 500 mA

fig. 3. Power supply for the five-band linear amplifier.

tuning and adjustment

The amplifier should be tuned to achieve maximum output at a given plate-current level while insuring that grid current does not exceed the normal operating value. When you have determined the normal settings for the tuning controls, it is easy to use these as starting points when changing bands. The amount of grid current at a given amount of plate current is determined by the setting of the loading control. With light loading, grid current will be excessive, while with heavy loading, grid current will be small, and efficiency will be poor.

There is some problem in determining just what the grid current should be under normal conditions. This depends upon the percentage of grid intercept. The manufacturer says that this is nominally 10%, but may be somewhat lower, or as high as 15%. This means that the grid current could run a bit higher than that listed on the manufacturers data sheet, and still not be excessive.

The best way to determine the correct grid current for a given power input is to check power output. My linear delivers 600 watts output on all bands except ten meters where it is a bit less. Allowing more than minimum grid current to achieve this power level does not result in more output, so this minimum grid current is my operating value. It is, of course, desirable to use a scope when operating and tuning up on ssb.

heat tests

Heat tests were run to determine the temperatures that the tubes and heat sinks reach during normal operation. The

temperatures were checked with temperature-sensitive compound sticks known as *Tempilstiks.** The heat sinks were allowed to stabilize with just the filaments running before the heat-test runs were started. Static dissipation tests were made by reducing the bias on the tubes until each tube was dissipating 200 watts.

The anode temperature (on the side of the tube away from the sink) reached 200° C in 8 to 10 minutes. There was a 50° C drop from the tubes to the heat sinks. The tubes must not be operated above 250° C, so a limit of 200° C under normal operating conditions is reasonable.

Tube temperature seemed to slowly increase beyond 200° C. This indicates that key-down operation for periods longer than ten minutes would require some additional cooling measures, such as a thermostat on the heat sink to turn on a fan. The thermal link blocks used on my tubes are 3/8-inch thick, but thinner blocks are available for improved heat transfer. However, thinner blocks add more stray capacitance to the plate circuit.

Since the heat sinks are attached to the chassis, it receives conducted heat. Therefore it is wise to avoid mounting parts that could be damaged by the elevated temperatures close to the heat sinks. The hottest spot on the chassis is in the area around the tubes.

It was noted while running the static heat tests that one tube drew about 10 mA more than the other, so it was running much hotter. Diodes were added as shown in **fig. 4** to place more bias on the higher current tube. The reverse diode provides a current path during the negative half of the rf cycle. This simple addition seemed to achieve balance between the two tubes.

Tube operating temperatures were also checked during operation. Under extended operating times, such as a CW ragchew, tube temperatures did not reach 200° C. Likewise, ssb operation at 1500 watts PEP did not result in excessive tube temperatures. Even hot and heavy contest operation doesn't overheat the tubes.

operation

The linear should be placed so air can freely enter and leave the cabinet since cooling relies primarily on air moving by convection upward through the cooling fins. The cabinet should be mounted on feet, ¾ inch or higher, to allow free entry of air into the bottom of the cabinet.

The thermal links should be checked from time to time to insure that they have not changed position. It might be possible to place a ridge of solder or epoxy on the heat sinks around the location of the blocks to prevent creeping.

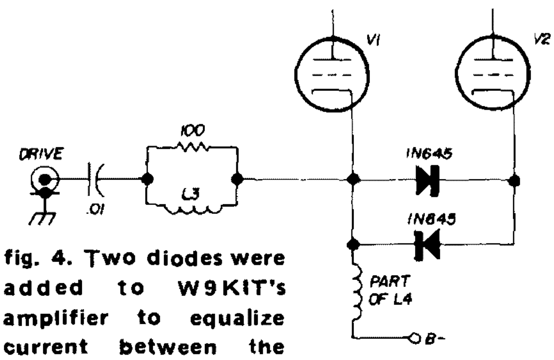


fig. 4. Two diodes were added to W9KIT's amplifier to equalize current between the two final-amplifier tubes (see text).

I have found this linear amplifier to be quite stable. The only problem I had was my homebrew exciter taking off on 80 meters. This was due to a poor ground connection between the exciter and the linear, and was cured with a ground strap between the units. I am well satisfied with the amplifier, and find that it fulfills all my operating requirements.

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**Tempilstiks* are manufactured by the Tempil Company, 132 West 22nd Street, New York, New York 10011.

crystal-controlled AFSK generator

Using one oscillator with a variable frequency divider, allows crystal-controlled AFSK without envelope aberrations

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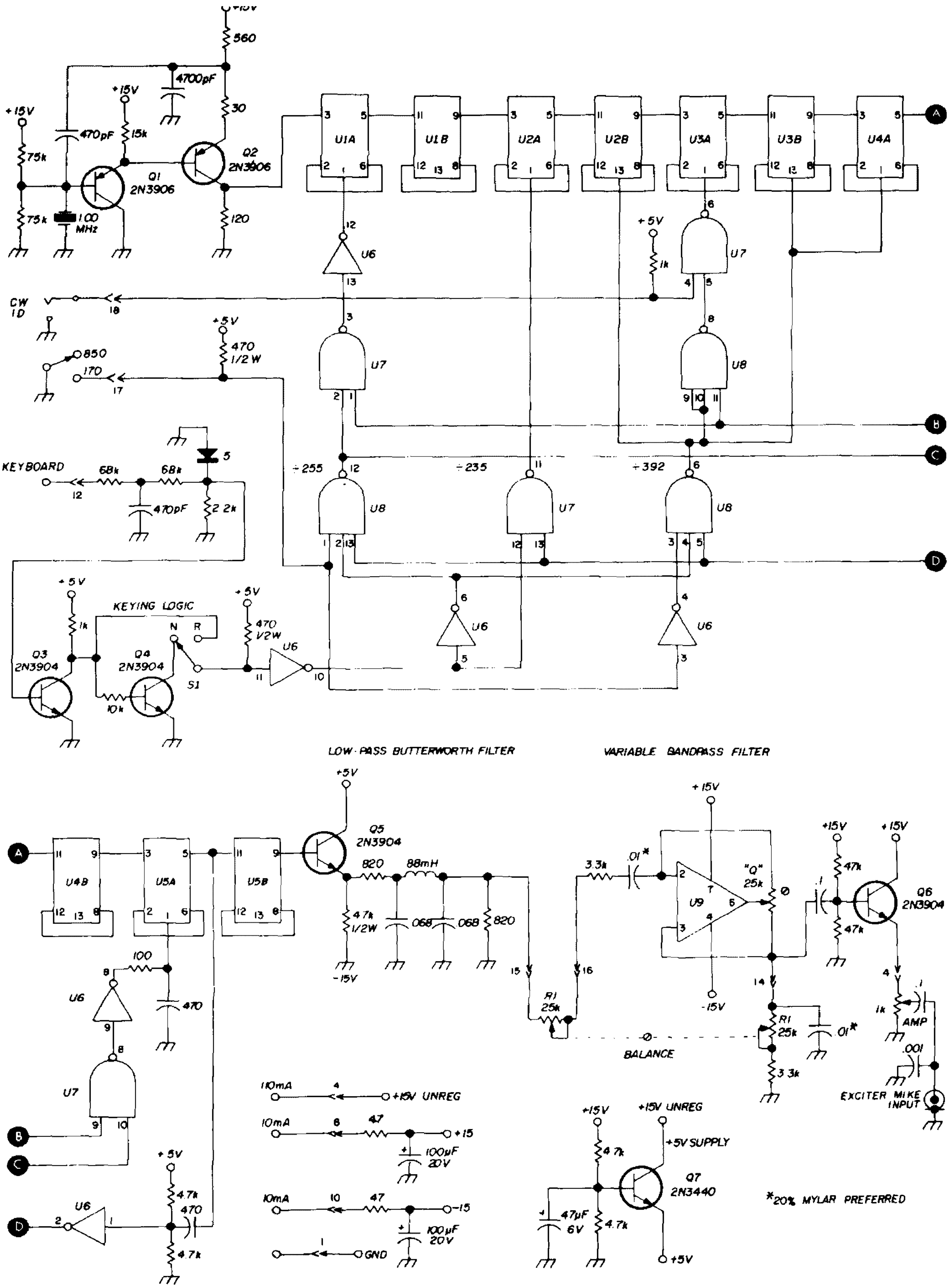
Several crystal-controlled AFSK circuits have been described using two separate oscillators and a common digital dividing circuit.[1] The circuit presented here is unique in that it uses a single oscillator and varies the division factor of the counter circuit. This has the inherent advantage that transitions between tones always occur at zero crossings, thus eliminating envelope aberrations often generated when switching between two non-synchronous tones.

I used a 1-MHz crystal because these crystals are readily available on the surplus market. Dividing 1 MHz by 2.125 kHz yields a ratio of 470.59. Since the counter can only divide by an integer I use a compromise ratio of 470, resulting in a output tone of 2.1277 kHz; 2.7 Hz higher than the standard mark frequency. Starting with a 100-MHz *oscillator and* dividing by 4,706 would give a 2.12504-kHz output, but the cost of ICs operating over 10 MHz and the problems of 100-MHz circuitry would not justify the added accuracy.

I used a standard ripple counter for *the frequency-divider circuit* (see **fig. 1**). These counters give an output with a non-symmetrical period when dividing by ratios other than two, four, eight and so forth. Since a non-symmetrical period represents even harmonics, the filtering requirements are eased by designing the basic counter to yield twice the desired frequency and then dividing again by two in a simple toggle circuit.

The output at pin 5 of U5 is differentiated through the RC network at pin 2 of U6 providing a *clear* pulse for the appropriate stages to provide the different ratios. For instance, when 2.127 kHz is required each time pin 5 of U5 goes negative, a pulse is coupled through the NAND gate logic scheme to the 1, 4, 16 and 256 stages. The counter *will start* each cycle with a count of



Q7	2N3440, TO-5 case, 1W	U6	SN7404 Hex inverter
R1	dual 25k potentiometer, single shaft	U7	SN7400 Quad 2-input NAND gate
U1, U2, U3, U4, U5	SN7474 Dual D-type edge-triggered flip-flop	U8	SN7410 Triple 3-input NAND gate
		U9	μA741 Operational amplifier

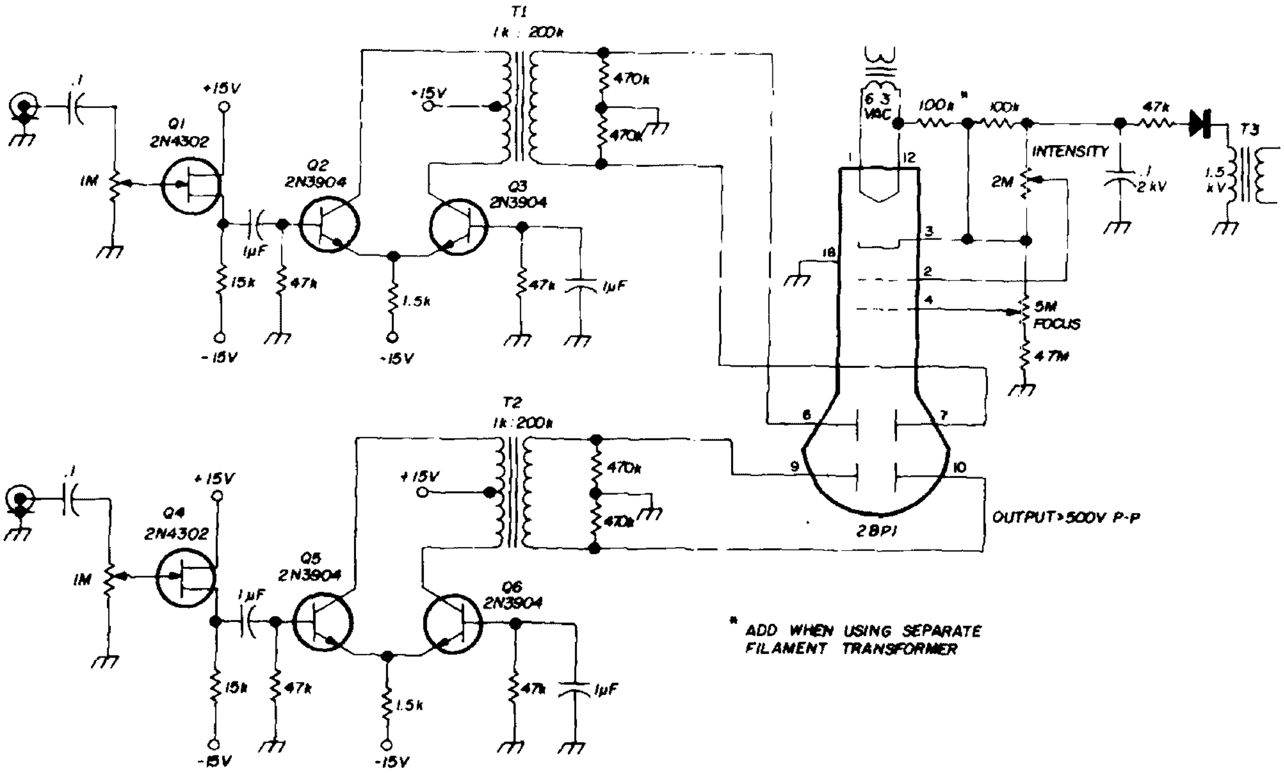


fig. 2. Transistorized, transformer-coupled deflection amplifier. The coupling transformers, T1 and T2, are inexpensive imported audio interstage types. T3 is a high-voltage scope transformer.

1+4+16+256=277, thus pin 5 of U5 will go negative at an actual count of 511-277=234. The output of U5 will be a non-symmetrical square wave with a frequency of 4.154 kHz providing a 2.127 symmetrical square wave at pin 9 of U6. Since the *clear* pulse is derived from the 256 stage, an RC network is placed ahead of the 256 clear input to insure enough delay for proper self-clearing action.

When a keying transition appears at the input (the network connected to the base of Q3), the counter ratio cannot change until the cycle in progress is completed.

The output of the counter is fed to a 3-pole Butterworth low-pass filter with a -3 dB point of 2.5 kHz. The low-pass filter is followed by an active variable

fig. 1. (opposite page) Crystal controlled AFSK generator. On every IC, U1 through U8, pin 14 is connected to +5 V and pin 7 is grounded. Divider resistors are chosen to give 5V at the emitter of Q7. Unless labeled, all resistors are quarter watt.

band-pass filter.[2] Tuning this filter between the mark and space tones provides amplitude compensation to counteract unequal attenuation of the two tones in the low-pass filter and the filter in the ssb exciter.

construction

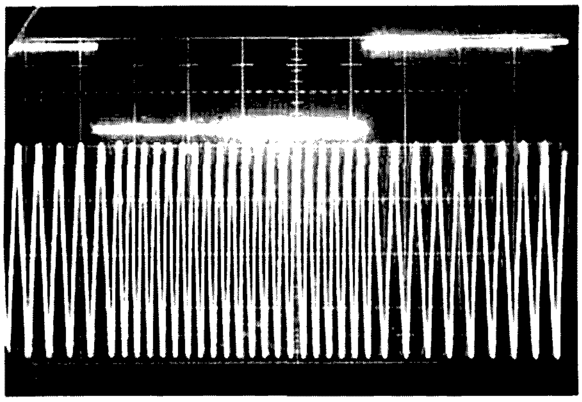
The entire circuit was built on a 22-pin plug-in Vectorbord. I used a BC-929 cabinet to enclose the AFSK circuit as well as a version of the ST-6 demodulator.[3] The ST-6 ±15 V supply provides enough current for the AFSK, ST-6 and a transistorized transformer-coupled deflection amplifier (see fig. 2) that drives the original BC-929 3-inch CRT.

The circuit is fairly complex for the average amateur with nine ICs and seven transistors. The new Motorola frequency-synthesizer ICs (MC4318) would simplify the design, but at least two would be required at $7.50 each versus less than $10 for all of the TTL ICs through surplus outlets. The total cost of components through surplus houses is around $25. The 5-volt supply is marginal and

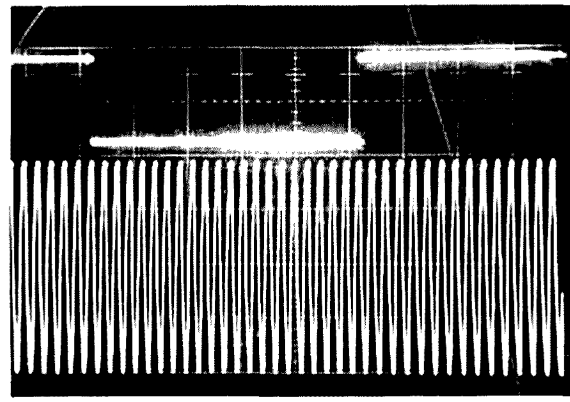
was used for economy. An independent rectifier running from a 6-Vac filament transformer with a Zener diode controlling Q7 would be a definite improvement.

lator with a single switch.

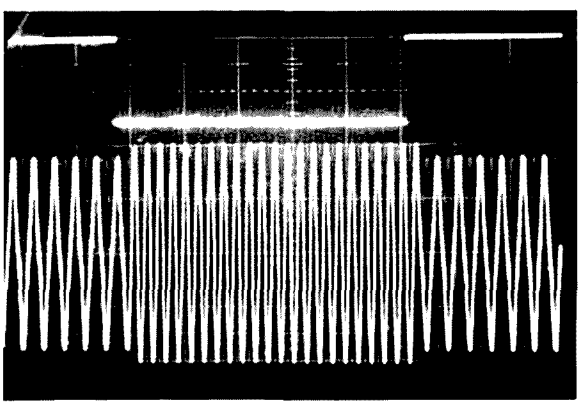
The low-pass filter does not remove the third harmonic of the 1.275 kHz tone completely, and this is compounded when the band-pass filter is aligned to



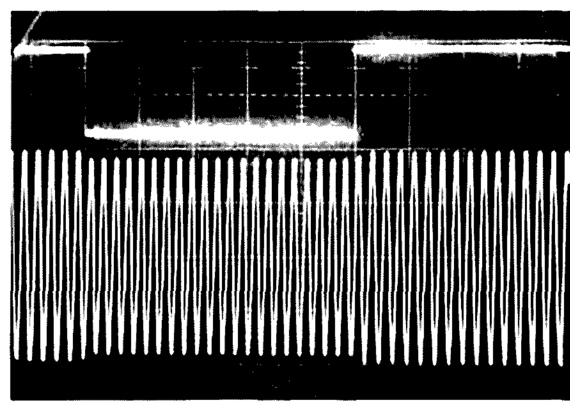
Wide shift adjusted for equal tone amplitude.



Narrow shift adjusted for equal tone amplitude.



Wide shift adjusted to boost high-frequency tone.



Narrow shift adjusted to boost low-frequency tone.

fig. 3. Performance of the AFSK generator. An electronic pulse generator was used to key the AFSK unit for these photos. All photographs were taken from a Tektronix type 547/1A4 50 MHz oscilloscope.

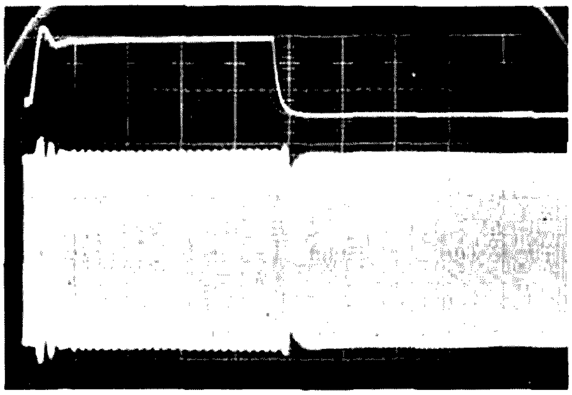
summary and criticism

The AFSK unit has been used to drive a HeathKit HW-100 for several months with very satisfactory results. While the output tones are not exact, they are more accurate and stable than typical LC or RC oscillators. For normal operation, I use upper sideband and I adjust my ST-6 tuned circuits to match the tones I use: mark–2125, space–1275, for 850 Hz shift; mark–2125, space–1955, for 170-Hz shift. The CW identification shift is 132 Hz. Since the change between narrow and wide shift requires only a ground closure, it is simple to control the AFSK shift and the shift of the demodu- boost the 2.125 kHz tone. An additional 88 mH toroid would provide a 5-pole filter which would reduce the third harmonic, but the ssb exciter filter seems to prevent radiation of the third harmonic.

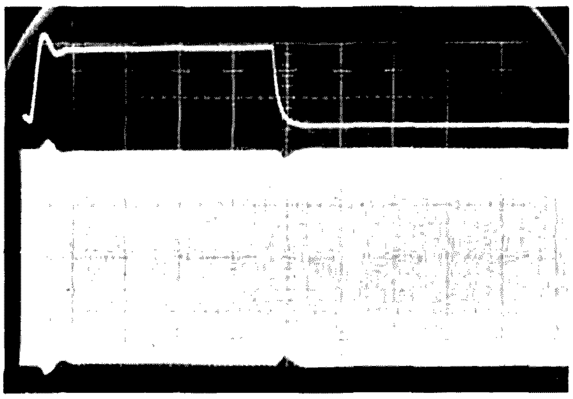
While the audio envelope does not have any aberrations and tone transitions occur at zero crossings, there is some ripple on the rf envelope during transitions. This is probably due to attenuation of audio sidebands in the sharp ssb filter.

The speed of transitions between tones could be slowed, using an RC oscillator with a damping network in the frequency determining voltage loop for instance,

thus reducing the audio sidebands and the rf envelope aberrations. This would increase the rise and fall times of the demodulated waveform, thus reducing the effective width of the mark and space



Rf envelope using wide shift.



Rf envelope using narrow shift.

fig. 4. Rf envelope pictures taken with HW-100 driving a Cantenna, 75 W at 3.6 MHz.

pulses, resulting in less immunity to noise, particularly on 100-wpm signals.

Thanks to W7GNI for many helpful circuit design suggestions, and to K7TBQ and WB6BZW/7 for help in evaluating the circuit.

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resistance-capacitance oscillators

Replacing the large inductors of the traditional L-C circuit with resistors and capacitors for today's miniature audio oscillators

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The trend to microcircuits and its resultant reduction in size of electronic equipment has been accompanied by a move to eliminate inductors. This is not only because inductors (at least those with henries of inductance) are large and heavy, but because the inductor is one component that will apparently not be put on a silicon chip. (At vhf and uhf where very small inductances are useful, inductances *can* and *are* being made using etched circuit and metal-deposited-on-substrate techniques.)

Because of the engineering drive to get rid of the iron, new ideas for circuits using resistances and capacitances instead of inductances and capacitances are rapidly coming to the fore. One of the areas where R-C replaces L-C is in filtering. The techniques for systematic multiple-pole low-pass, high-pass, and bandpass filters using only resistances, capacitances and IC operational amplifiers are pretty well established.[1] Tuned transformers associated with ratio detectors and discriminators are also on their way out, having been designed around by IC phase-locked loops and other IC circuits for fm demodulation.

Another area wherein R-C circuits are widely used is oscillators; unlike the circuit designs mentioned above, R-C oscillators are not new technology. R-C oscillators have been with us for decades, at least for audio frequencies.[2] The

phase-shift oscillator, bridged-tee oscillator, twin-tee oscillator, and Wien Bridge oscillator (with vacuum tubes) are good examples. Circuits of these are shown in **fig. 1** as they were originally used, with vacuum tubes as the gain blocks. Only the equivalent ac circuits are shown; that is, no blocking capacitors or biasing resistors are included.

basic circuits

The circuits presented in **fig. 1** have only one thing in common—all are R-C types. These circuits differ significantly in details. The phase-shift oscillator and the twin tee oscillator shown in **figs. 1A** and **1B** have no amplitude control system, and require their R-C networks to shift 180° at the oscillation frequency. The bridged-tee oscillator shown in **fig. 1C,** however, uses the R-C network as a null *in the negative feedback path to increase* the loop gain at the oscillation frequency. Positive feedback is *broadband* and provided by the resistance of I1 in **fig. 1C.**[3] The phase shift of the bridged-tee network is 0° at the null frequency. Like the bridged-tee oscillator, the Wien Bridge oscillator has an R-C network which provides 0° phase shift at the oscillation frequency. The R-C network, however, provides a peak instead of a null at the oscillation frequency, and so is placed in the *positive* feedback loop.

In **figs. 1A** and **1B** no attempt is made to automatically control the amplitude of oscillation. For this reason, these two circuits will generally produce a somewhat distorted sine-wave output.

The circuits of **figs. 1C** and **1D,** however, have a non-linear resistance (I1) which increases its resistance with an

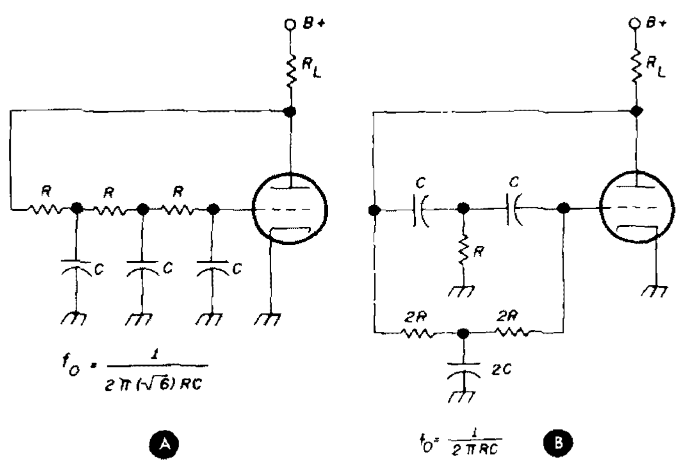
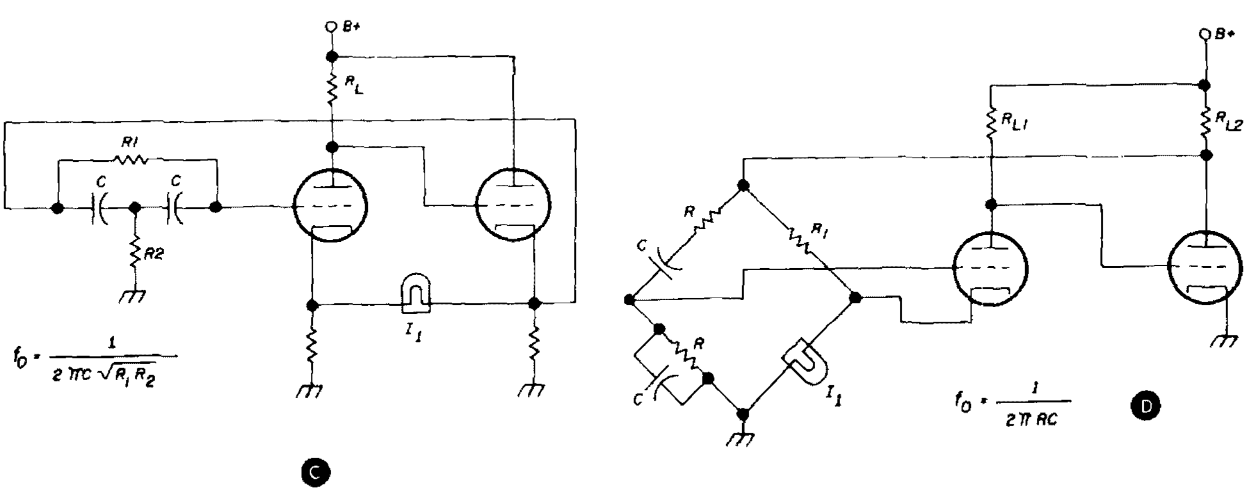


fig. 1. Some basic R-C oscillators: phase-shift (A), twin-tee (B), bridged-tee (C) and Wien Bridge (D). The bridged-tee and the Wien Bridge both contain non-linear resistances which result in a more pure sine wave than that produced by the phase-shift or twin-tee oscillators shown.



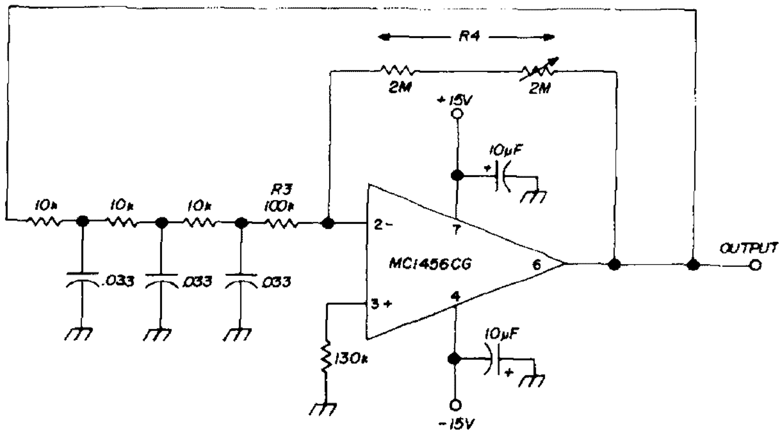


fig. 2 Phase-shift oscillator using monolithic IC op amp. The phase-shift section is composed of the three 10k resistors and the three 0.033 μF capacitors at left. All pin numbers refer to the TO-5 package.

increase in output level. The series position of I1 in **fig. 1C** assures that if the output level increases the effect will be to *decrease* the positive feedback factor. The parallel position of I1 in **fig. 1D** has the effect of increasing the negative-feedback factor upon an increase in output. In both cases the effect of I1, in increasing resistance with an increase in level, is to restore the operating condition that existed before the change.

The two oscillator circuits, with a nonlinear resistance for feedback control, produce very nearly sinusoidal waveforms, since the automatic feature holds them in the linear operating region. For this reason, these circuits have been widely used as laboratory audio generators; the famous Hewlett Packard 200 series is based on the Wien-Bridge circuit. Similarly, the bridged-tee is used in the Heathkit IG72 audio generator. It is also feasible, however, to apply amplitude control to the circuits of **figs. 1A** or **1B.**

op amps

Let us now redraw the four R-C oscillators using operational amplifiers as the gain-blocks. These are shown in **figs. 2, 3, 4** and **5**. The resistor R3 establishes the input impedance into which the R-C network must operate. Also, in *all* of these circuits, the ratio R4/R3 determines the closed-loop gain of the amplifier.

In the basic phase-shift oscillator, there are three R-C sections, each with 60° of phase shift at the operating frequency. These phase shifts all add up to 180°, causing the input and output to be 360° out of phase so that oscillation will start if the gain of the amplifier is adequate.

The twin-tee oscillator operates in much the same fashion as does the phase-shift oscillator. At the frequency of oscillation, there is 180° phase shift through the R-C network. The twin-tee network will be remembered by some readers as a notch network, which was often used in a-m radios to suppress the 10 kHz interstation whistle in early hi-fi tuners. In the oscillator use, however, it never quite operates in the notch, or no positive feedback could occur.

In the bridged-tee oscillator of **fig. 4**, the amplifier has an all-pass network (resistive voltage divider) in the positive-feedback path; so that except for the presence of negative feedback, it could oscillate at *any* frequency. However, the bridge-tee network is a *null* network, and so gives *minimum* negative feedback at its *null* frequency. This minimum negative feedback at the null frequency means that the amplifier has its maximum gain at that frequency, and so that's where the oscillations occur.

The Wien Bridge as shown in **fig. 5** is quite different in that the R-C network does not control the negative feedback but rather the *positive* feedback. The negative feedback is controlled by R3 and R4.

All the circuits of **figs. 2, 3, 4** and **5**

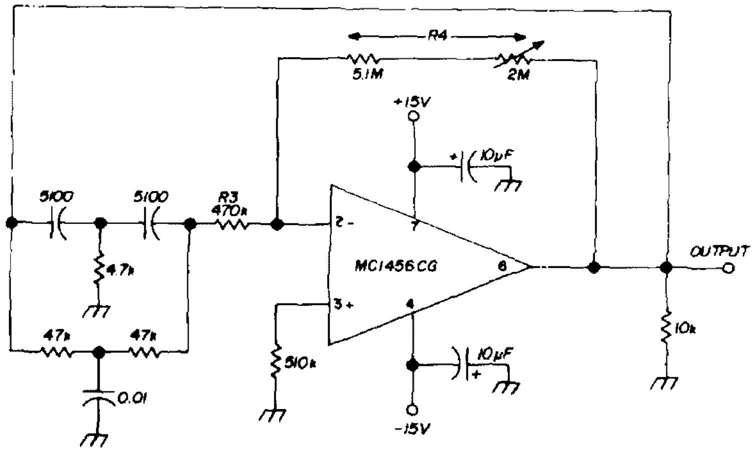


fig. 3. Twin-tee oscillator using an IC op amp.

use relatively low impedance values in their R-C frequency-controlling networks because of the low input impedances of the monolithic IC op amps they use. By using op amps with higher input impedances (such as the fet-input types), it is possible to make the C values smaller. This seemingly tiny advantage is really quite significant; it allows some important circuit possibilities: use of 15 to 468 pF dual- or triple-section variable capacitors for tuning, use of varactors for tuning, and use of on-chip capacitors. A few of the things that can then be constructed are: a capacitively-tuned continuously variable R-C oscillator, a voltage-variable R-C oscillator (for generating fm) and an R-C oscillator entirely built on a silicon chip.

variations

There are also some important variations on the basic oscillators that are worth looking at. The phase-shift oscillator, for instance, need not have the simple form of **fig. 1**, wherein all the Rs and Cs have equal values. A useful variation is that of tapering the R-C network.[4] This method uses a larger value of R and a smaller C in each successive R-C section. In this manner, the C of the first section is not loaded by the R of the following section (see **fig. 6A**). If R3 is much larger than R2, R2 is much larger than R1, C1 is much larger than C2, and C2 is much larger than C3, true tapering results. By "much larger than," I usually mean ten times as large. However, five times as

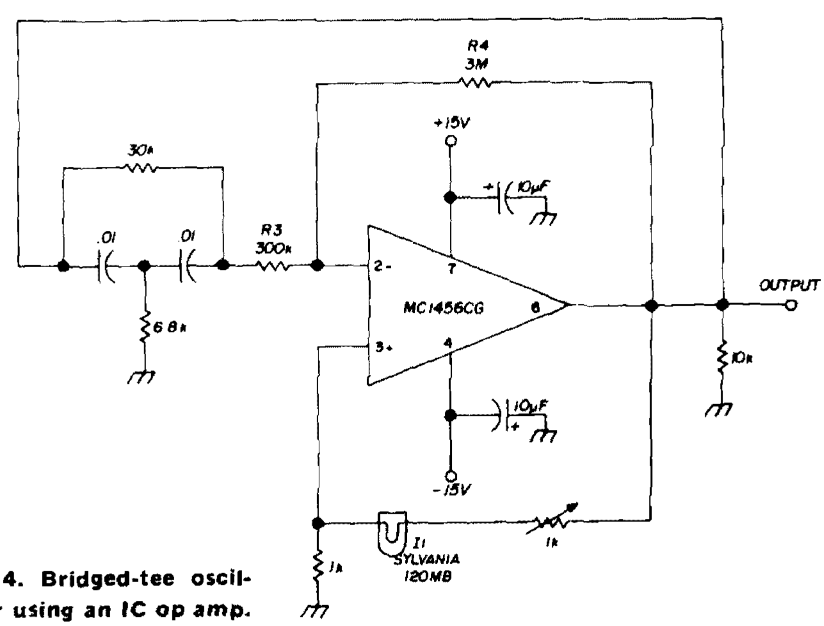


fig. 4. Bridged-tee oscillator using an IC op amp.

large is adequate for tapering in a phase-shift oscillator.

Fig. 6B shows another variation of the phase-shift oscillator network, using four sections. This variation can be extended to as many sections as desired.

Finally in **fig. 6C**, the Rs and Cs are exchanged in the circuit, and the phase-shift oscillator still operates. This works because it doesn't matter whether there is 180° lead or lag for oscillations to take place. This network may have some advantages in some circuits since it makes coupling capacitors and the amplifier input resistor part of the network. A concrete example of this is shown in **fig. 7**.

Fig. 7 also combines the techniques of *multiple sections (more than 3), and* tapering, to show that these several variations are all compatible. Tapering and increasing the number of sections both have the effect of reducing *the loss* in the R-C network, allowing us to use less amplifier voltage gain to sustain oscillation. A simple, three-section, equal R and C phase-shift oscillator takes a minimum voltage gain of 29 to make it oscillate. A four-section tapered network may require a voltage gain of less than 5 to oscillate.

A rather interesting *modification of* the phase-shift oscillator uses a *distributed* R-C line as the network.[5] This

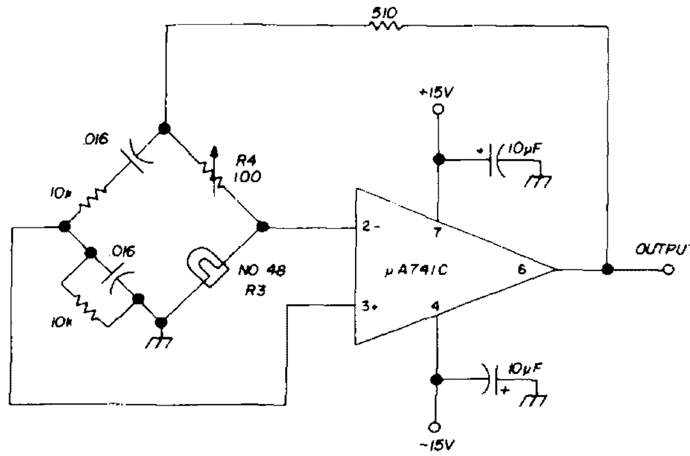


fig. 5. Wien-Bridge oscillator using an IC op amp. All pin connections refer to the TO5 case.

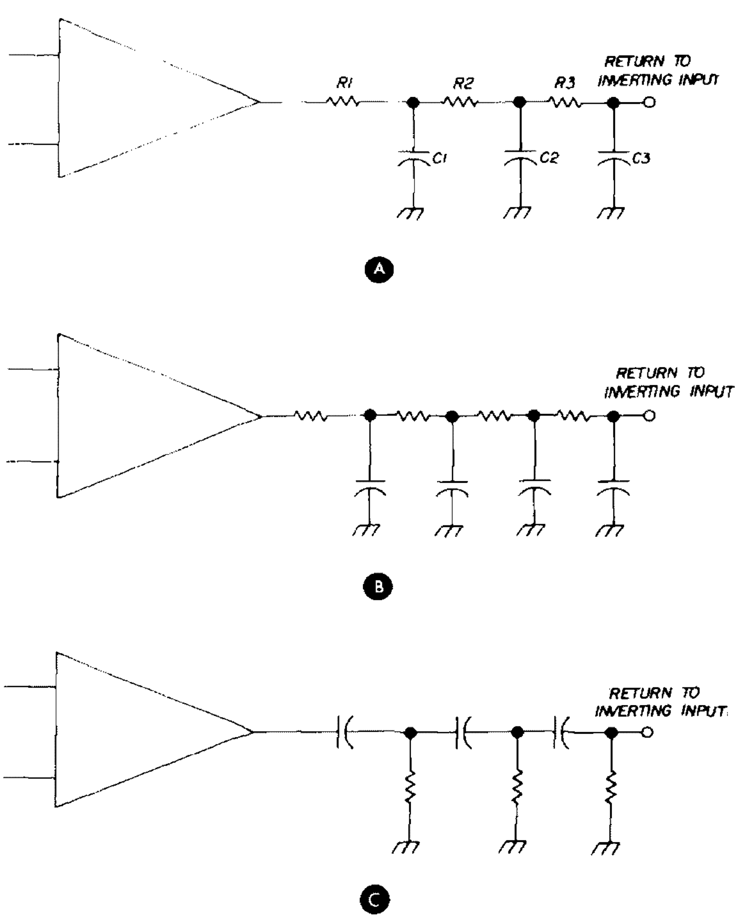


fig. 6. Variations of the basic oscillator. The tapered network is shown in A. C1 is much greater than C2 which is larger than C3. Similarly, R3 is greater that R2 which is greater than R1. B shows an extra section added to the R-C network. Any number of extra sections could be added. The usual location of the resistors and the capacitors has been switched in the oscillator in C.

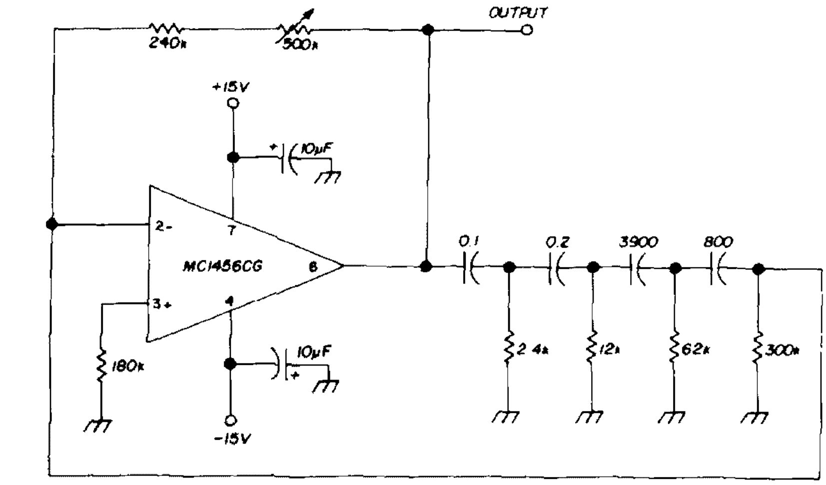


fig. 7. Phase-shift oscillator with series C and shunt R, multiple sections (more than three), and tapering.

could be thought of as a piece of coax cable with a high-resistance center conductor. It is then effectively a multisection R-C network with its shunt capacity to ground continuously in effect along *the length of the line. Such a* distributed R-C line phase-shift oscillator was actually built and tested as shown in **fig. 8**. The R-C line was made up of a deposited-carbon resistor (40K) of the older non-encapsulated style. A layer of ordinary kitchen-type aluminum foil was wrapped around its body and taped in place. This foil served as the outer conductor of the line, and the paint on the resistor formed the dielectric. The circuit oscillated at about 200 kHz, so it was necessary to use an IC op amp which had a high slew rate; I chose the Signetics NE531T because its slew rate is uniform at about 30 V/µsec regardless of the closed-loop gain. The NE531T is not internally compensated, however, and requires a single external capacitor for this purpose (100 pF between pins 8 and 6).

There are some other interesting ways of modifying the phase-shift oscillator. Combining tapering with a distributed line suggests itself, but is a bit difficult to implement using standard components. However, this is the sort of thing that may be very easy to do on a silicon chip in ICs of the future.

Wien Bridge variations

The Wien-Bridge oscillator has been well worked over; many variations have appeared. Early efforts to make a Wien-Bridge oscillator using bipolar transistors were inevitably tuned by a dual variable resistor, or used switched Rs and Cs. This,

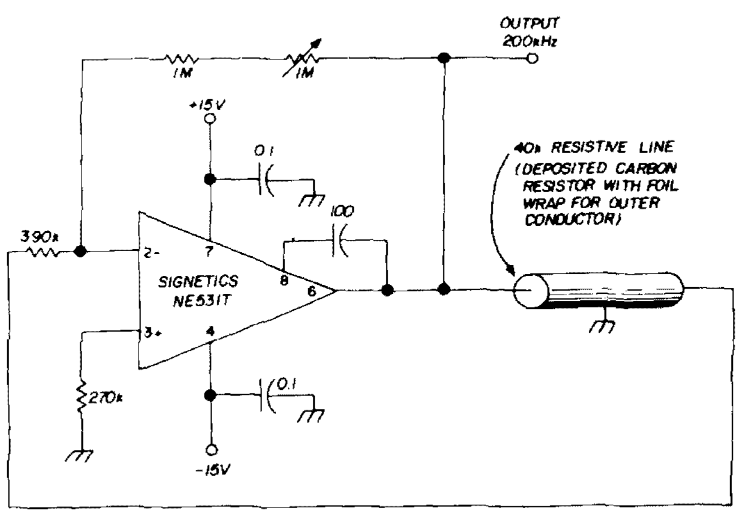


fig. 8. Phase-shift oscillator using distributed R-C line as the phase-shift network.

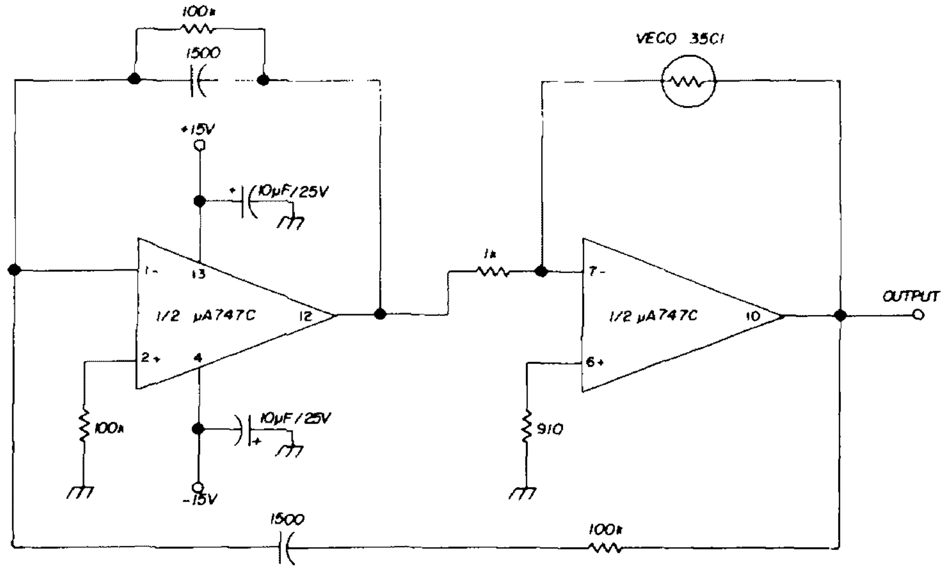


fig. 9. Baxandall version of the Wien Bridge. Notice the use of a thermistor as the nonlinear resistor in the feedback control circuit.

of course, was simply due to the low input impedance of the base of a bipolar transistor. An interesting variation of the Wien Bridge is the Baxandall modified Wien Bridge circuit as shown in fig. 9.[6] The main reason for using the Baxandall version of the Wien Bridge is to prevent the op amp input impedance from loading the parallel R-C branch of the bridge. A practical circuit before op amps used four 2N404s.[7]

Fig. 9 also demonstrates the use of a thermistor as the nonlinear resistor in the feedback control. Unlike the lamp bulb, the thermistor *decreases* resistance as it is heated; so it is placed in the position shown to increase negative feedback as it heats. This type of feedback control element is used in at least one laboratory audio oscillator, the General Radio 1311A.[8]

It is also possible to rectify the output of the oscillator and apply the filtered dc to some form of agc element. The agc element can be any one of a number of devices such as a diode, transistor or fet.[9]

An fet can also be used as a voltage-controlled resistance, and so is sometimes used in the negative-feedback voltage divider of the Wien Bridge oscil-

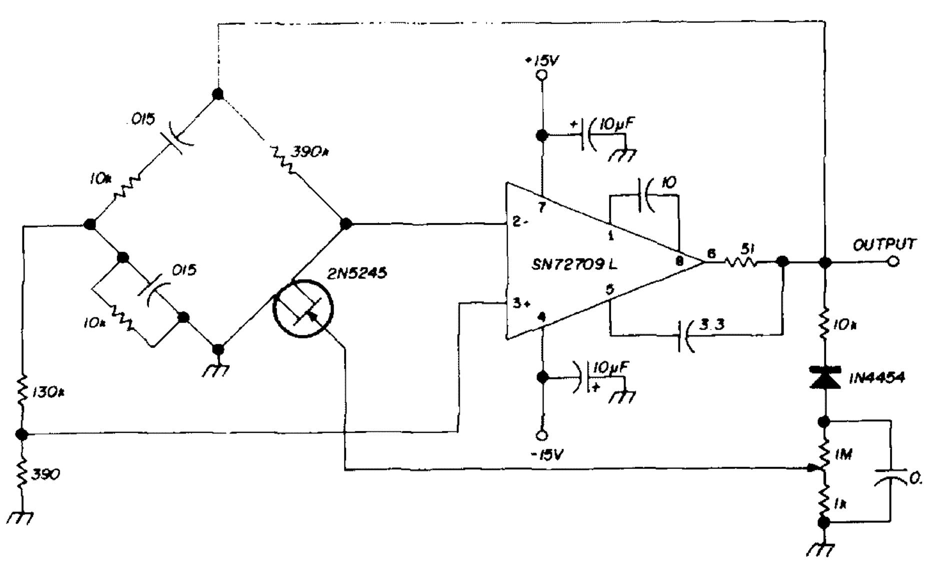


fig. 10. Wien Bridge using an op amp for gain and a fet for feedback control.

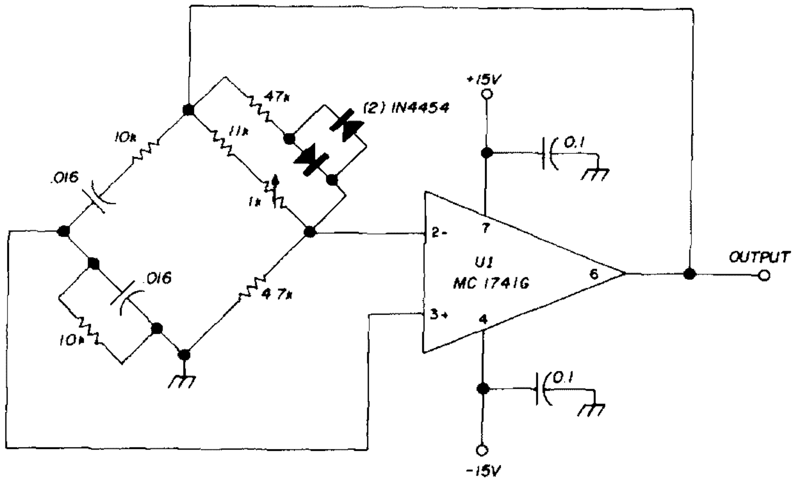


fig. 11. Wien Bridge oscillator using diodes for non-linear control element.

lator.[11,12] The source and drain terminals of the fet are treated as the two terminals of a resistor, and the control voltage is applied between source and gate. The fet used as a voltage-variable resistor should be operated with zero dc from source to drain and with very small ac voltage across it. Special fets characterized as voltage-controlled resistors are Siliconix.[13] An actual example of a Wien Bridge oscillator using an fet for feedback-control and an IC op amp as the gain stage is shown in **fig. 10.**[12]

There is one version of the Wien-Bridge oscillator that has no time constant inherent in the negative feedback control element.[14] This circuit is shown in **fig. 11**. As the nonlinear element is a pair of back-to-back diodes essentially operating instantaneously, there is no time constant to them. In fact, the negative feedback portion is very much like an operational voltage clipper. The 47-kilohm resistor in series with the diode pair and the fact that the diodes go into conduction symmetrically, however, allow this oscillator to produce fairly pure sine-wave output.

Of course, there are many other types of R-C oscillators besides the four basic circuits covered here. The ones which produce non-sinusoidal waveforms, such as astable multi-vibrators, would provide enough subject material for another article. Even in the restricted area of sinusoidal oscillators there are a couple more types that have not been covered here. These types are rarely used, and in some the reasons for their invention (low mu limitations of early tubes) no longer exist.

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ham radio

converting the
Motorola Dispatcher
to 12 v operation

Using an inexpensive homemade toroid to convert these surplus bargains from 6 V to 12 V operation

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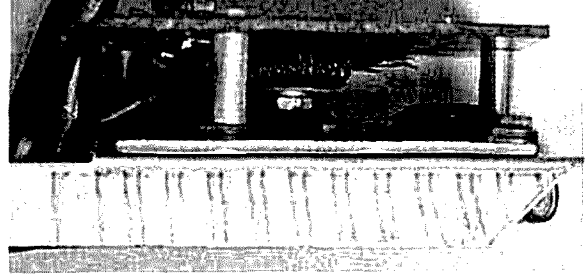
There are a lot of Motorola Dispatcher 6- and 2-meter transceivers just waiting to be put into service by amateurs. These partially trasistorized radios started becoming available several years ago as police departments decided to replace their old equipment. Sometimes called motorcycle radios, the units are reliable, small, easy to tune to amateur frequencies and are relatively inexpensive. They are priced as low as $30.00, including accessories, mainly because of their not-so-popular 6 V power supply. Dispatchers with 12 V supplies are available, but not as readily or inexpensively.

The 6 V supplies can be modified for 12 V operation the conventional way, but the expense can easily equal the cost of the entire radio. The 12 V transformer alone is priced in the tens of dollars. For this reason, many hams resort to a series requlator to drop their 12 V system to 6 V for the radio. But this method is terribly inefficient, and the cost is about

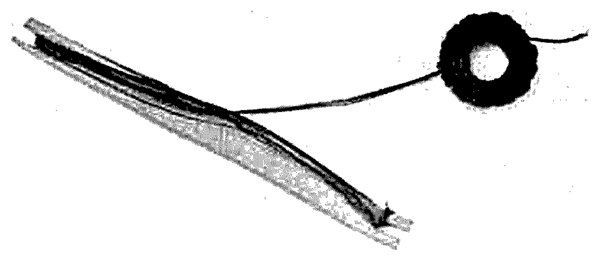


The new toroid is mounted on the phenolic platform when working with a newer model dispatcher.

the same as the method described here.*

theory

The principle behind the conversion is simple. In the original circuit, the transformer primary is a 6-0-6 V winding, fig. 1A. The switching transistors alternately connect the 6 V battery between the center tap and one 6 V winding, and the center tap and the other 6 V winding, figs. 1B and 1C. The alternating field then induces various voltages into the other windings of the transformer. While Q1 and Q2 are conducting, fig. 1B, winding W2 is open and has a voltage induced in it. Likewise, while Q3 and Q4 are conducting, fig. 1C, W1 is open and has a



The home-wound toroid and the cardboard bobbin.

aiding, the two 6 V potentials add to 12 V between A and C. During one half of the cycle, A is positive 12 V with respect to C, and during the other half, A is negative 12 V with respect to C. So, as it turns out, switching the 6 V battery in

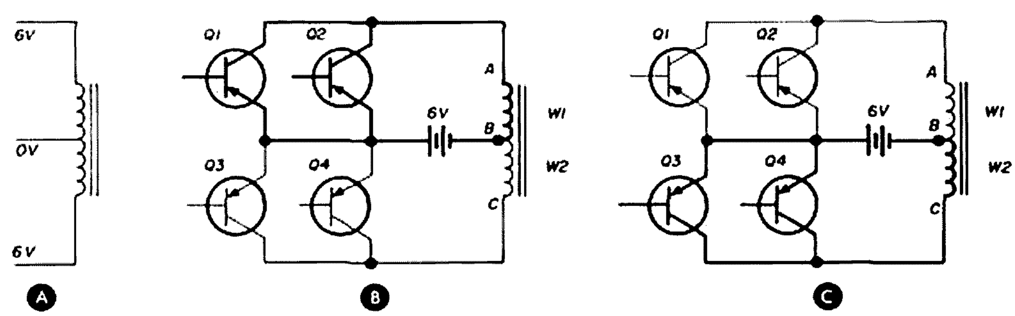


fig. 1. Original circuit for the Dispatcher power supply. B and C show current flow during alternate half cycles.

voltage induced in it. Since W1 contains the same *number* of turns as W2, when 6V is switched across either winding, then 6 V is induced into the other. Furthermore, because the windings are in series

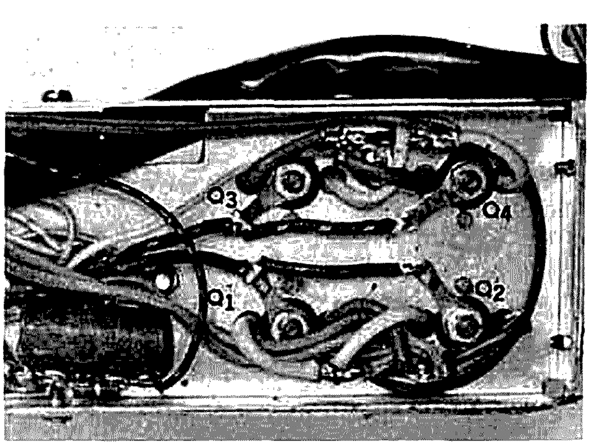


Photo of the power supply shows transistor location and chassis orientation for best parts access.

this fashion is equivalent to switching a 12 V battery across the entire winding (A-to-C). This is precisely what is accomplished by the modification described here.

The original supply consists of four transistors wired in push-pull-parallel. Their configuration is changed to that of a quasi-complimentary bridge for the 12 V supply. During one half of the cycle, Q1 and Q4 conduct, connecting the 12 V battery between A and C with positive on A (fig. 2A). During the other half of the cycle, the polarity is reversed by the action of Q1 and Q4 turning off as Q2 and Q3 turn on (fig. 2B).

Drive is obtained by the addition of a

*A complete parts kit, including a wound toroid, is available for $6.50 postpaid from John Darjany, WB6HXU, 2347 Angela, Suite 4, Pomona, California 91766.

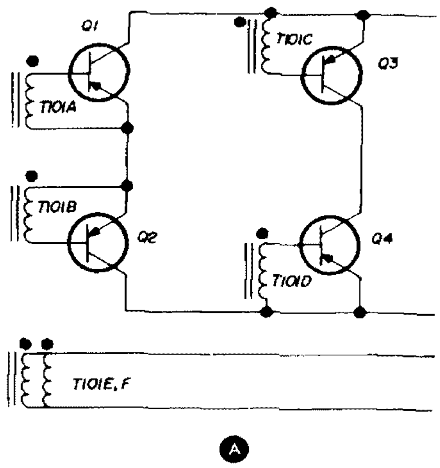
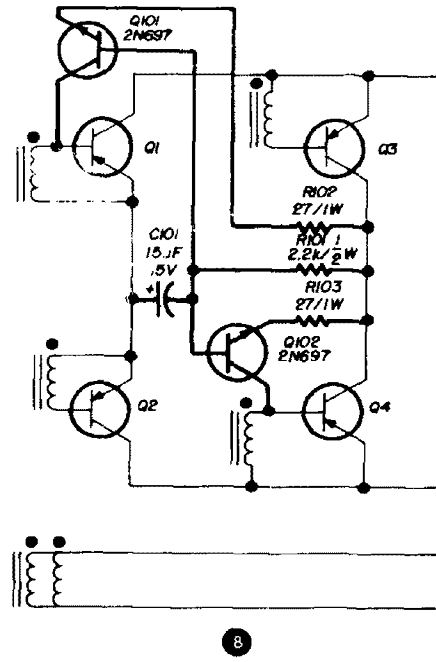


fig. 3. Drive circuit of the modified power supply uses resistance of the toroid windings for base-current limiting and balancing (A). The starting circuit is shown in B.



small home-wound toroidal transformer which utilizes the dc resistance of its windings for base-current limiting and balancing (fig. 3A). The start circuit, fig. 3B, generates a pulse rather than a continuous level so that receive current drain is held to a minimum (about 500 mA). The original 6 V circuit may be compared with the complete modified 12 V circuit for further clarification (figs. 4 and 5).

modification procedure

The first step in the modification is to

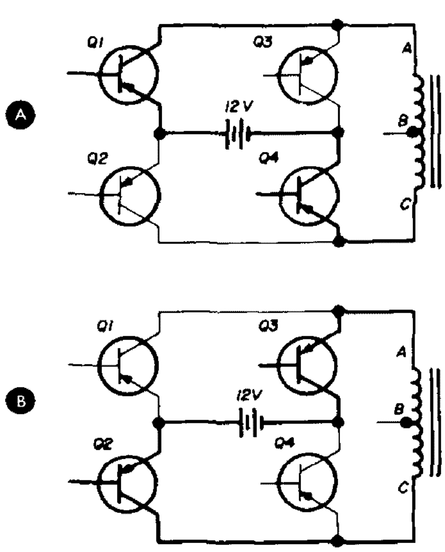


fig. 2. Basic operation of the circuit modified for 12 V operation. Again, A and B show current flow during alternate half cycles.

wind the driver transformer. Cut a bobbin out of stiff cardboard and wind around it about 9 feet of six strands of number 30 enameled copper wire. Then, using the bobbin as a sewing needle, thread 80 turns of the six-strand bunch onto a Magnetics 55206-A2 toroid core, leaving about six inches of wire at each end.* Set the transformer aside until later.

Remove the power supply chassis from the bottom compartment and position it upside down in front of you, oriented as in the photographs. Note that there are several versions of this supply, and some may appear somewhat different than the ones shown. For example, one later version uses TO-3 transistors rather than TO-36 as pictured here.

Unsolder Q2 (see the photo for transistor locations) and remove it from the chassis. Install a mica washer with some thermal joint compound, reinstall and rewire Q2. Next, remove the jumper between the collector of Q1 and Q2. Remove the black and yellow wire from the transformer at the collector of Q3. Remove the emitter wires from Q3 and Q4 at the terminal strip. Connect the emitter of Q3 to the collector of Q1. Connect the emitter of Q4 to the collec-

*Magnetics Component Division, Butler, Pennsylvania 16001.

tor of Q2. Extend the black and yellow wire from the transformer with some number 16 wire, and connect to the collector of Q2. Remove all four base wires from the transistors at the terminal strips and remove the jumpers between adjacent terminals. Remove the yellow and white and the red and white wires coming from the transformer at the terminal strip. Connect the base wires of the four transistors individually to the four empty terminal lugs. Disconnect the black and brown wire from the transformer and tape it so that it will not short to something.

The phenolic mounting platform for the toroid is part of the newer supplies. Avoid short circuits by mounting it so that no sharp or hard edges or surfaces contact the wires. Connect a wire from the collectors of Q3 and Q4 to the negative supply (heavy green wires coming from terminals 2, 3, 4 and 5 on the power connector). If an earlier version supply is being converted, mount the toroid below the chassis, held in place by its leads and the green wires.

Remove the 330 ohm, ½ W resistor. Remove the white and brown wire from the transformer and tape the free end. Remove the positive supply wire from the 25 ohm, 10 W adjustable resistor and connect the resistor to the white and red wire from the transformer. Change the 220 ohm, ½ W resistor to a 220 ohm 1 W resistor.

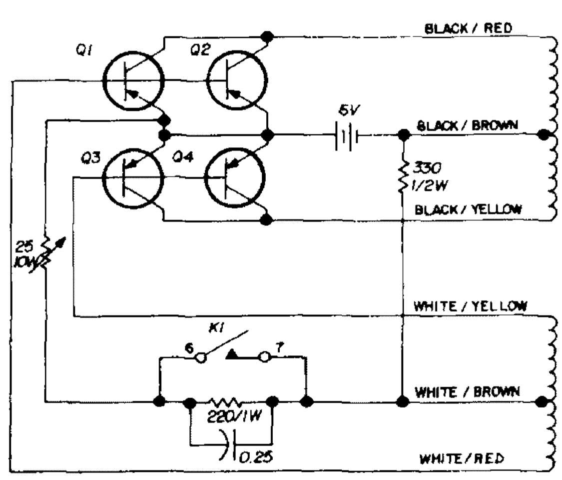


fig. 4. The original power supply circuit set up for 6 V operation.

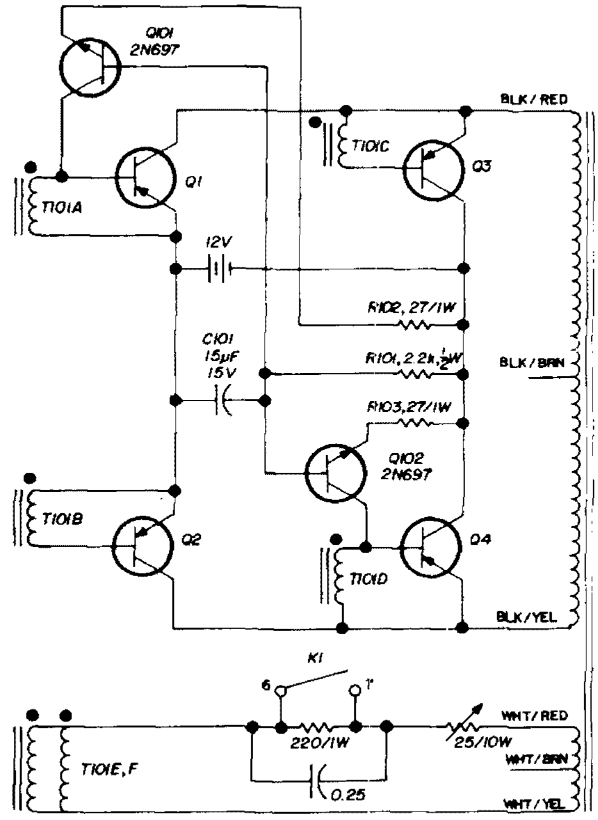


fig. 5. The completely modified power supply set up for 12 V operation. T101 is 80 turns, six strands no. 30 magnet wire on a Magnetics 55206-A2 toroid core. It is described in the text.

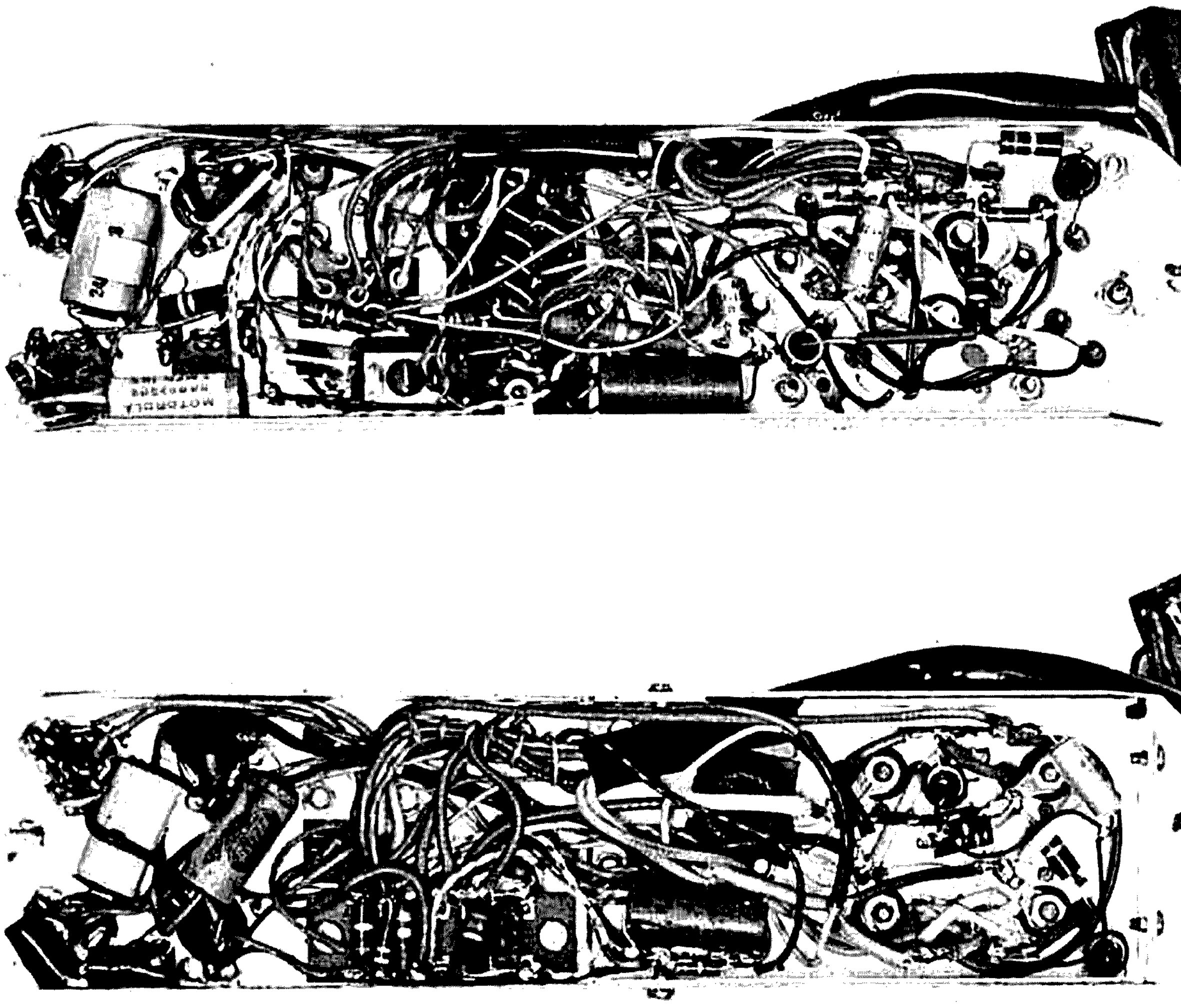
toroid wiring

For clarification, the six wires leaving the center of the toroid nearest the chassis will be called positive, and those leaving the center of the toroid away from the chassis will be called negative. This will help in connecting the windings in proper phases.

Connect any two positive toroid wires to the 220 ohm, 1 W resistor at the side not connected to the 25 ohm, 10 W adjustable resistor. Find the two corresponding negative wires with an ohmmeter, and connect them to the white and yellow wire coming from the original transformer. Connect any one remaining positive wire to the base of Q1, and its corresponding negative wire to the emitter of Q1.

Connect another positive wire to the base of Q4 and its corresponding negative wire to the emitter of Q4. Connect a remaining negative wire to the base of Q2

and its corresponding positive wire to the emitter of Q2. Connect the last negative wire to the base of Q3 and the last positive wire to the emitter of Q3. The wiring of the starting circuit is straightforward and can be done by following the diagram of fig. 3B and the photos.



Underchassis view of the completed modification. The later model Dispatchers is shown in A, while B shows the earlier model. Notice on the earlier units how the toroid is mounted under the chassis, held in place by the surrounding wires. Tape under the toroid protects the enameled wires from the steel chassis.

operation

The modification is now complete and should require no adjustments. If, however, the transmit voltages are low, careful adjustment of the 25 ohm, 10 W resistor should help. Of course, it is advisable to disconnect power while making adjustments.

Most of the parts used should be available in the average junk box. The 2N697 transistors can be replaced with any silicon npn switching transistor with similar characteristics. There are surplus toroids available which can replace the one called for – but the new, single-unit price is much less than a dollar, so surplus store hunting may not be worth the effort.

The results have been excellent. One unit has been in 12 V operation for nearly three years, and has not given any trouble at all. Another was completed along with this article, and looks as good as the first. Inquiries accompanied by a return envelope will be answered gladly.

ham radio

optimizing the superregenerative detector

Hangover and receiver radiation can be minimized with these circuit refinements

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The superregenerative detector, known since the early 1920's, has seen limited commercial application but has appealed to radio amateurs and radio-control enthusiasts. Few circuits are capable of providing such high gain from a minimum of conventional parts, not to mention additional advantages such as low power requirements and good high-frequency performance.

The superregen circuit has been described in previous issues of *ham radio*.[1,2] The problems of hangover (blocking that limits sensitivity) and receiver radiation were shown to be minimized by adding a diode across part of the tank inductance.[3]

While many subtleties are involved in circuit operation, I've found that two simple factors are outstanding in achieving optimum effeciency: time constant and applied bias, which influence the blocking-voltage waveform.

The superregenerative circuit is a blocking oscillator[4] operating at radio frequency. The circuit contains an RC network with a time constant long with respect to the natural frequency of the circuit. After a few cycles of oscillation, a reverse bias develops on the RC network. The reverse bias becomes sufficiently large to cause circuit losses to exceed circuit gain. Residual oscillation then decays, and the remaining dc bias bleeds off until the conduction point of the circuit active device is reached; then the process repeats. The process is illustrated in **fig. 1**, a typical blocking- oscillator waveform developed at the input.

The rate at which the circuit goes into and out of oscillation is the quench frequency, which is made to occur at a supersonic rate so as not to interfere with the received signal. Detection occurs because any rf signal (or other electrical disturbance) will cause the bias slope to encounter the conduction point ahead of its natural time, thus increasing the quench frequency. Since a higher quench frequency means a higher average current through the circuit, a replica of the signal modulation is available.

The key to successful operation of the circuit is in the quenching voltage waveform, which influences circuit performance in two major ways. It determines:

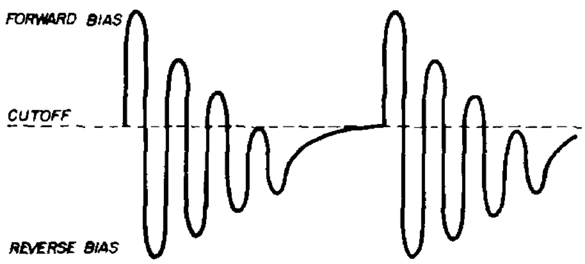


fig. 1. Typical blocking-oscillator wave-form developed at input of superregenerative circuit.

a. The amount of audio that will be developed from a given input signal.

b. The degree of regeneration a given signal will experience prior to the actual circuit oscillation burst.

It is this latter factor which controls circuit selectivity and sensitivity.

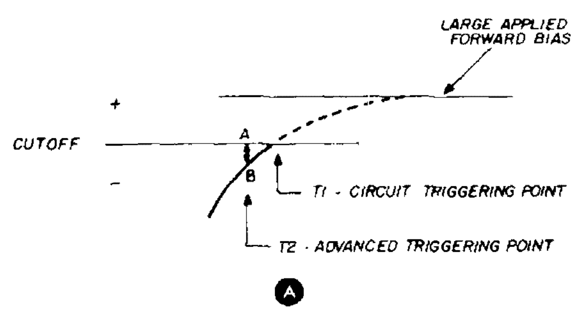
Consider now the effect of the quenching voltage waveform on circuit audio response, as illustrated in **fig. 2.** (In this figure the initial oscillation burst has been eliminated since it doesn't directly pertain to the following discussion.) In **fig. 2A** a relatively long time constant is used with a large value of forward bias. The forward bias increases the quench *frequency* by eliminating the shallow portion of the decay slope. The remaining portion of the slope crosses the circuit firing point (approximately equal to cutoff in the case of transistors) at a steep angle. If a signal having an instantaneous amplitude represented by the line A-B is applied to the slope, the circuit fires at T2 rather than at T1. The increase in *quench frequency* is small, resulting in a relatively low audio output.

The circuit conditions for **fig. 2B** are the same except that forward bias has been greatly reduced, preserving the shallow portion of the decay slope. If the same signal represented by line A-B is applied to this slope, a much greater change in quench frequency is produced, resulting in a much higher audio output. Unfortunately, the reduction in forward bias also results in a normal quench frequency well within the audio range, which destroys the usefulness of the circuit. What is important here is that (a) slope *shape* rather than quench frequency *per se* influences circuit efficiency, and (b) a shallow slope depends on the application of minimum forward bias.

Having established one condition for optimum performance, the second problem is to eliminate the low quench frequency. This condition can be achieved by reducing the value of the time-constant network: if the time-constant is made as short as possible, the net effect will be a usable, supersonic quench frequency with a shallow slope.

selectivity and sensitivity

The most significant advantage of the



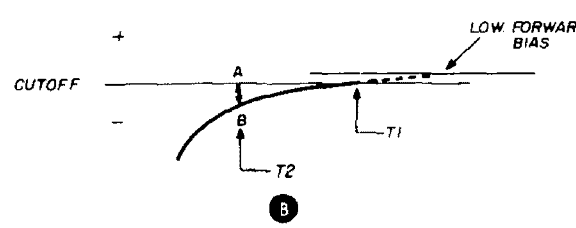


fig. 2. Triggering-point slope and quench frequency. In A a signal of amplitude A-B advances triggering time from T1 to T2. Increase in quench frequency is small since shallow portion of blocking waveform is lost due to large forward bias. Reduced forward bias, B, preserves shallow portion of slope and signal A-B causes a large change in quench frequency.

low *forward bias, short* time-constant combination is not the production of high amplitude audio but its influence with regard to circuit selectivity and

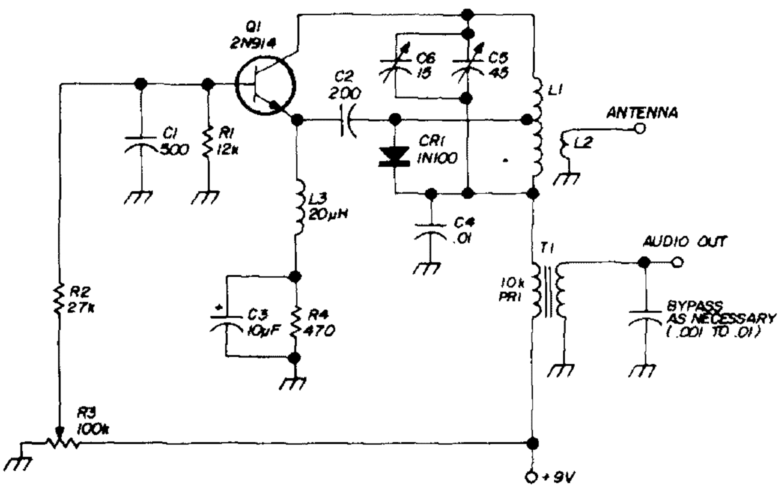
sensitivity, which is related to the amount of pure regeneration present in the circuit. It is well known that simple regenerative receivers achieve a Q multiplying effect by operating close to the unity gain or oscillation point.

The superregenerative circuit is also regenerative when the decay slope is near the unity-gain bias level. Here, the advantage of a shallow slope, as compared to a steep slope, is that the shallow slope permits the signal to dwell in the threshold region for a longer period. This action allows a greater regenerative buildup with a resultant increase in gain and decrease in bandwidth. Since circuit losses are compensated by regeneration, the Q of the tank circuit doesn't greatly influence selectivity as it is sometimes believed. In fact, an excessively high Q tank is not permissible in the circuit.

hangover

In the superregenerative circuit, it is essential that no oscillations remain in the tank circuit after oscillation has terminated. Such persistent oscillation, sometimes referred to as hangover, can create spurious responses that block detection by forcing the receiver to listen to its own residual rf signal.

Hangover, which results from the large reactances involved, prevents efficient operation in the lower frequency ranges. In this case efficient operation can be obtained only at excessively low quench frequencies, which allow time for natural tank damping. Furthermore, attenpts at constructing a truly narrowband circuit, such as using a quartz crystal in conjunction with the tank, are doomed to failure unless some method of damping is developed.



L1 14 turns no. 24 on 5/16" diameter form, tap 5 turns from ground end

L2 4 turns no. 24 around cold end of L1

fig. 3. 28 MHz superregenerator.

The performance of conventional superregen circuits can be improved by adding a damping diode across a portion of the tank inductance. The diode is connected so that it doesn't prevent positive feedback. The diode dissipates energy immediately after the circuit oscillation burst, thus hangover is eliminated. Since the barrier potential (approximately 0.2 volt in the case of germanium diodes) exceeds any normal signal input, no tank loading will occur. The advantages resulting from diode inclusion are:

1. Residual oscillations are eliminated from the sensitive portion of the decay slope.

2. A more highly regenerative (shallow) slope can be employed since the tendency of the circuit to lapse into cw oscillation will be reduced.

3. Receiver radiation is greatly reduced: Diode damping lowers the amplitude and shortens the duration of the radiated pulse. (Preliminary measurements indicated a 12:1 reduction of radiated noise in the case of the 28-MHz circuit described below.)

Two circuits are presented, which exemplify the superregen operational principles. The first (**fig. 3**) covers the 28-MHz band and tunes from approximately 21 to 40 MHz. The second circuit, **fig. 4**, covers the 144-MHz band.

L1 3 turns no. 19 on 5/16" diameter form, tap at 1 turn from ground end

L2 2 turns no. 19 around cold end of L1

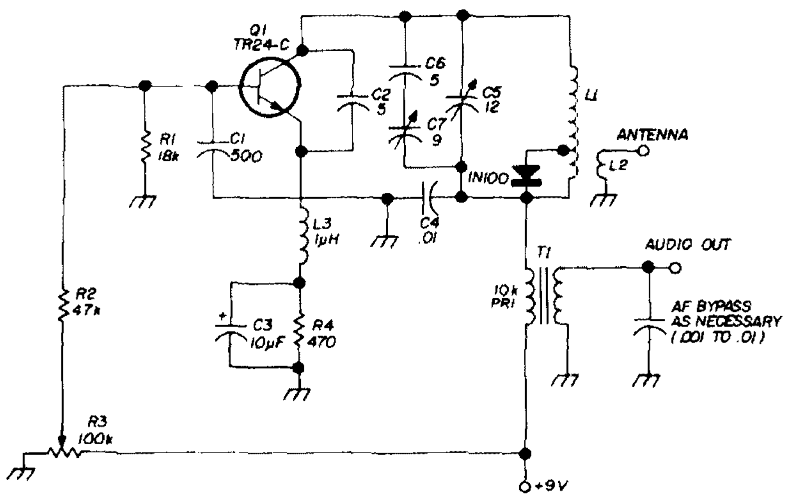


fig. 4. 144-MHz superregenerator.

The difference between these circuits in the use of a Colpitts oscillator in the 144-MHz range to compensate for shunt capacitance.

In both circuits, the time-constant network consists of R1, C1. As a general rule, the time constant of R1, C1 should be as short as possible, the limiting factor being the point at which the circuit tends to lapse into cw oscillation. In this regard, one variable requiring compensation will be the beta spread within given transistor types, with high-gain units requiring a longer time constant. Here, the best method of adjustment consisted of holding C1 at 500 pF and adjusting the value of R1. When making this change isolating resistor R2 should have a value at least twice that of R1. Potentiometer R3 controls the forward bias and acts as a quench-frequency control. The most sensitive setting will correspond to the minimum bias consistent with a usable quench frequency – usually in the order of 20 kHz. Occasionally, this sensitive setting of R3 will result in oscillator pulling from strong local signals. If this occurs, the control should be advanced to a slightly less sensitive setting.

A simple resonant circuit and diode detector were placed near the antenna of the 28-MHz receiver to detect receiver radiation. The output was fed to a scope and displayed as a vertical line. When the diode was inserted, the radiation-pulse amplitude fell to 20 percent of its former value. The pulse width (scope sweep on) also decreased to about 70 percent of its former value.

The net effect on a nearby linearized receiver (no avc; no rf overload) was to reduce the audio noise voltage measured at the receiver output to 1/12th of its former value. This reduction amounts to about 21 dB. The 144-MHz circuit demonstrated the same effect; thus the principles appear to apply to every superregenerative circuit, allowing operation beyond the point at which a conventional circuit would be limited.

references

1. R. Brown, K2ZSQ, "No-radiation, No-hangover 28-MHz Superregen Receiver," *ham radio,* November, 1968, p. 70.

2. Atuyuki Iwakami, JA1BHG, "Improved Superregenerative Receiver," *ham radio,* December, 1970, pp. 48-49.

3. Charles L. Ring, "Diode Improves Performance in Superregenerative Circuit," *Electronic Design,* March 14, 1968, p. 230.

4. Hank Olson, W6GXN, "Enter the Strange but Useful Blocking Oscillator," *ham radio,* April, 1969, pp. 45-49.

ham radio

cooled preamplifier

for
vhf-uhf reception

An experiment into lowering solid-state device noise figure by cooling the unit hundreds of degrees

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One of the ingredients necessary for weak-signal reception at uhf is a receiver with the lowest possible noise figure. A lower noise rf amplifier in your frontend can make weak, fading stations easily readable as well as bring new signals up, out of the noise.

As a result of improved solid-state devices, noise figures have dropped substantially in the last few years. Today, a common-gate fet amplifier will readily provide a noise figure of 1.5 dB or better at 144 MHz. With this kind of performance from such a simple circuit, any improvement would be more or less academic. At 432 MHz, where low noise figures really begin to pay off, you can use a transistor with a possible noise figure of 1 dB. Fets are slightly inferior at this frequency but can give a 2 dB noise figure. Performance at 1296 MHz is similar, with state-of-the-art noise figures being approximately 2 dB for transistors and 3 dB for fets. In spite of all this, the parametric amplifier is still king. At 432 MHz, 0.5 dB can be achieved, and 1 dB is possible up to 2.3 GHz. However, the paramp represents a rather formidable project; at least it does to all of us who have never built one.

theory

In order to see how much these low noise figures can help your receiving setup, we'll review briefly a few expressions. These are:

$$(S/N)_o \propto 1/T_n \qquad (1)$$

$$T_n = T_s + T_e \qquad (2)$$

Where T_s = source noise temperature
T_e = effective receiver noise temperature
T_n = system noise temperature
$(S/N)_o$ = output signal to noise ratio

The equations tell us that the output signal-to-noise ratio is inversely proportional to the system noise temperature. Further, noise figure is related to receiver noise temperature by

$$T_e = (F - 1)290°K \qquad (3)$$

Noise performance, therefore, is specified by either. It should be pointed out that F *in* **equation 3** *is the ratio* noise figure, not in dB. These are related by **equation 4.**

$$F_{db} = 10 \log F \qquad (4)$$

Now, T_S depends on several factors, but if everything is right, can be as low as 200°K at 144 MHz and 20°K at 432 MHz for a moonbounce link. If you're not familiar with this noise figure business, try reading K6MIO's article, and then plug in a few numbers and see what happens.[1] It shouldn't take long to convince yourself of the value of a low noise figure.

The approach to a low-noise amplifier used here is based on the fact that most noise added to the amplified input by an fet is thermal noise and proportional to the physical temperature of the device. Thus we should find that **equation 5** is true.

$$T_e \propto T_{device} \qquad (5)$$

test procedure

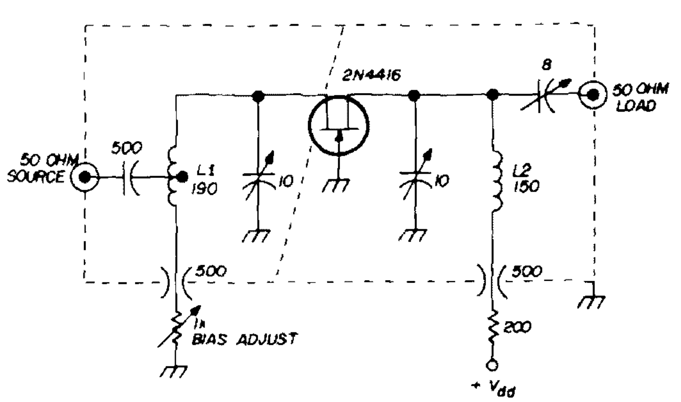
To test this hypothesis, I built the circuit of **fig. 1**. You may wonder why the common-gate configuration was used instead of the common-source. This is simply because a common-gate amplifier will give just as low a noise figure as a common-source circuit. This is so even though the *device* noise figure is lower in common-source. The input tuned circuit has a higher loaded Q and hence more loss than the corresponding circuit in the common-gate amplifier. Also remember that as we get rid of device noise, the lossy components preceeding the fet become important. Therefore, it is desirable to use the common-gate configuration with its low-Q input.

Before cooling the amplifier, I adjusted the circuit for best noise figure at room temperature. I adjusted the supply voltage, drain current, input coil tap, output coupling capacitor and input and output tuning capacitors. I found that the output coupling capacitor had little effect on noise figure. I finally set it at about 2 pF. The position of the input-coil tap was not critical, but a minimum was obtained with it about 30% up from ground.

Input tuning was broad. Again, however, there was no question that a dip in noise figure did exist. With the output tuning I obtained best noise figure at approximately the position of the capacitor that gave maximum gain. This point was, however, with the circuit tuned slightly on the high side of the signal frequency.

The results of varying the supply voltage and drain current differ somewhat from what is usually recommended as optimum. Most amplifiers use a supply voltage of 10 to 15 Vdc and a drain current of approximately 5 mA. I found with this amplifier that the best noise figure occured at $V_{gs} = 0$ and $V_{ds} = 5$ Vdc.

To clarify this somewhat, if $V_{gs} = 0$, a



L1 6 turns, no. 16 enamelled copper wire, 3/8-inch inner diameter, input tapped 1¾ turns from ground

L2 5 turns, no. 16 enamelled copper wire, 3/8-inch inner diameter

fig. 1. 144 MHz common-gate fet test circuit. The inductances are in nanohenries.

table 1. Tabulated results of cooling a 144 MHz common-gate, fet preamplifier.

device temperature T_{device} (°K)	noise figure F (dB)	drain current I_{ds} (mA)	supply voltage V_{dd} (Vdc)
300	1.5	7.9	6
200	1.2	9.8	8
77	0.8	16.	12

rather high drain current results, which requires that the supply voltage be low so that power dissipation and hence noise figure stay down. Turning things around, if we start with a high supply voltage, then drain current must be low for the same reason. Even with I_{ds} = 7.9 mA (V_{gs} = 0) and V_{ds} = 5 Vdc, you would see an increase in noise figure as the device rose to operating temperature.

All the adjustments mentioned lowered the initial noise figure from 2.0 dB to 1.5 dB, and although there was no guarantee that the amplifier would remain optimized upon cooling, this seemed the best way to start.

results

The results of cooling the preamp are tabulated in table 1 and the data plotted in **fig. 2**. The 200°K temperature was achieved by holding a small cube of dry ice directly on the case of the 2N4416 with a pair of small tweezers. Some slight retuning was necessary to minimize the noise figure. Liquid nitrogen at 77°K was used to get the lowest point in **fig. 2**. The amplifier was carefully lowered about a half inch into a Dewar containing the liquid nitrogen.

The three noise-figure readings are plotted in **fig. 2** and may be connected by a straight line. This shows that the noise temperature of the amplifier is proportional to the physical temperature of the fet. It is seen that at T_{device} = 0, T_e is greater than zero. This is due to circuit and transmission-line loss preceeding the fet and noise added by receiver stages following the cooled preamp. In this case the preamp was followed by two common-gate amplifiers and a transistor mixer into the noise-figure meter.

Because of the relatively high antenna noise temperature at 144 MHz, the low receiver noise temperature at 144 MHz, the low receiver noise temperature obtained cannot be fully appreciated in actual use. However, at 432 MHz this is not so, and if similar device behavior at this frequency is assumed, then a noise figure of 1.2 dB should be possible upon cooling with liquid nitrogen. This number is obtained by drawing a graph similar to **fig. 2** and taking the noise figure at 300°K to be 2.5 dB. While this is good, a parametric amplifier and *some* transistor amplifiers can do better.

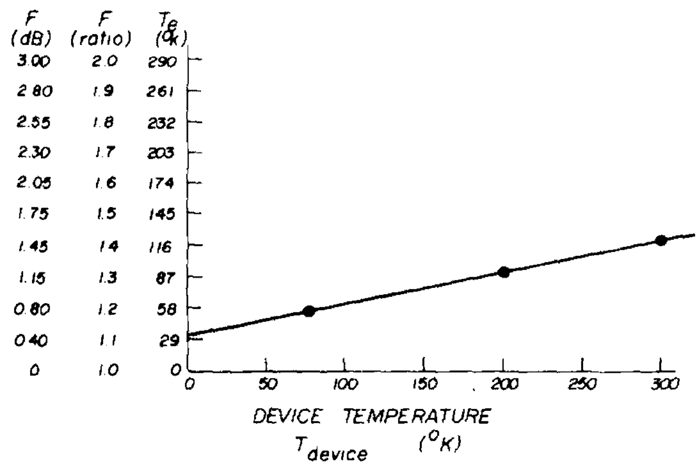


fig. 2. Noise figure and noise temperature are plotted against device temperature for the 144-MHz preamplifier shown in fig. 1. The results are tabulated in table 1.

The possibility of improvement with further experimentation should be evident. Barring some physical change in the operation of the device, cooling with liquid helium at about 4°K would give an extremely low noise temperature. Also, the possibility of operation above 1 GHz should not be overlooked since low-noise *microwave fets may* be just around the corner. I'd be very interested in hearing results from anyone who might try something along these lines.

reference

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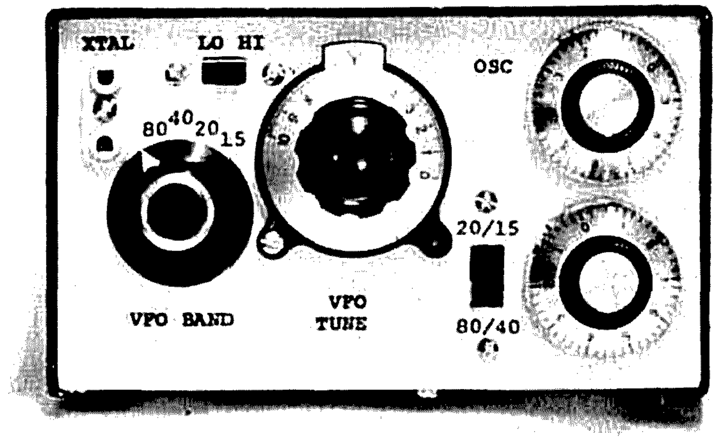
ham radio



a multiband fet vfo QRPP transmitter

For the real low-power newcomer this rig covers the 80- through 15-meter bands with an inexpensive rf module and a versatile vfo

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Operating with less than five watts of power is lots of fun, but the newcomer to the sport of QRPP often faces many unexpected problems. This is particularly true when the newcomer chooses to build his own solid-state gear.

First, most of the published designs for solid-state gear are exclusively for crystal control. This is a serious drawback in actual QRPP operation where the weaker QRPP signal is often smothered by the higher-powered interference. The ability to change frequency to dodge interference is essential to successful operation. Also, the weak QRPP signal is less likely to work stations by calling CQ than by answering another station's CQ on the other station's frequency.

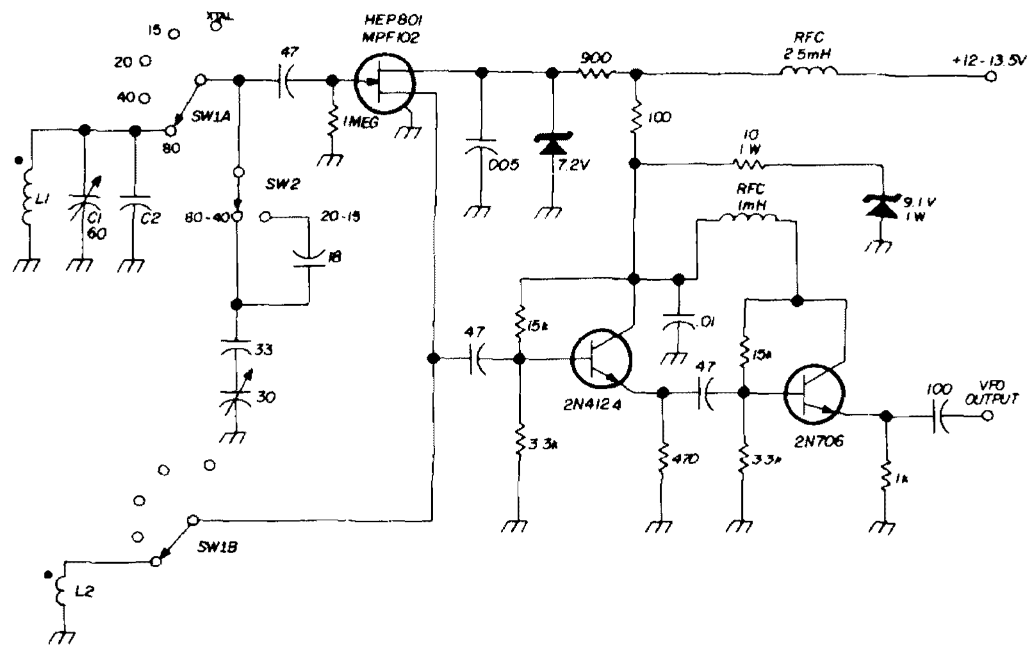
Second, most solid-state gear is designed for the 80- or 40-meter bands or vhf, totally overlooking the bands which provide the greatest opportunity for consistent success — 20 through 10 meters. This is unfortunate because the propagation and lower atmospheric loss factors of these bands foster low-power operation. For example, during the past year I have worked about 40 states including Hawaii and the Commonwealth of Puerto Rico

on 20 and 15 meters while using only a simple antenna and an output never exceeding 120 milliwatts. Any QRPPer seriously interested in DX must be able to work these bands.

Third, the QRPP neophyte building his first solid-state rig often is baffled by the

factory wired and tested, has bandswitching capability for 80 through 15 meters, is able to put out between 0.6 and 1.4 watts and has performed well in thousands of installations.

The multiband vfo (fig. 1) uses the HEP-801, available at most radio supply



C1	60 pF trimmer (Elmenco 404)
S1	2 pole 5 position rotary switch
S2	single pole double throw switch
S3	double pole double throw switch or preferably two single pole double throw switches
S4	double pole double throw switch

fig. 1. Schematic of the multiband fet vfo. All capacitors given in picofarads are silver-mica dipped; all bypass capacitors are ceramic.

problems of duplicating rf circuits. This is certainly true of any attempt to construct a multiband driver-final circuit.

The solid-state transmitter described here provides the interested amateur with an economical, practical and dependable rig. Frequency flexibility is achieved with a bandswitching fet vfo that is stable and chirp free. I took advantage of a readily available Ten-Tec TX-1 driver/final module* to make it easy for the beginner to construct a solid-state transmitter without encountering the difficulties associated with rf power-stage construction. In addition, the TX-1 is

*Available from Ten-Tec, Inc., Highway 411 East, Sevierville, Tennessee 37862 for $7.95 postpaid.

stores; an MPF-102, or its recent derivatives – the 2N5668, 2N5669 or 2N5670 – will also work. I achieved isolation and buffering through two emitter-follower stages and using isolating

table 1. Tank circuit values for the fet vfo.

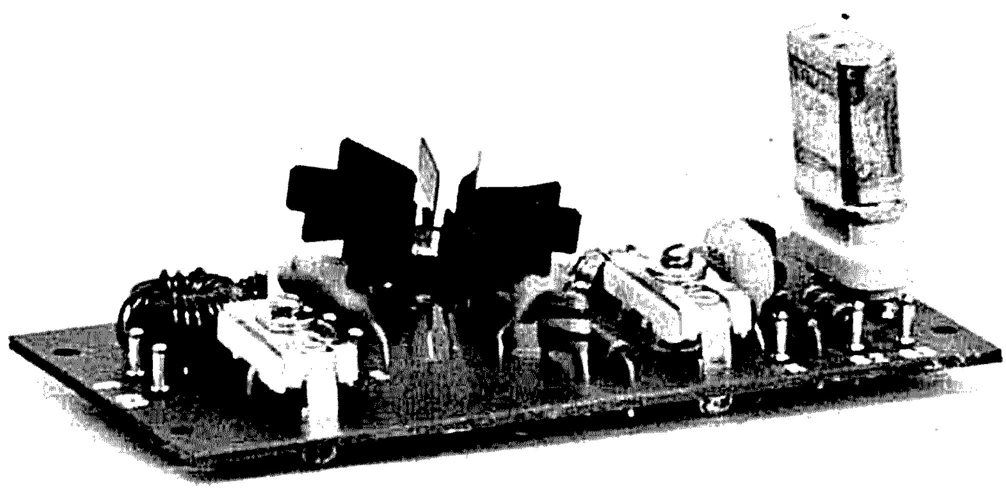
band	L1	L2	C2	coverage
80	50 t. no. 28	17 t. no. 24	47 pF	225 kHz
40	19 t. no. 28	9 t. no. 24	180 pF	110 kHz
20	8 t. no. 24	4 t. no. 24	150 pF	120 kHz
15	7 t. no. 24	3.5 t. no. 24	100 pF	340 kHz

All coils are wound on Amidon T-50-2 cores except for the 15 meter coil which is wound on a T-38-2 core. In all cases, L2 is wound over L1 over the full diameter of the core. Both windings start at the dots shown in fig. 1 and are wound in the same direction! All wire is enameled copper.

rf chokes in the B+ leads to the oscillator and final buffer. Zener-diode regulation of the oscillator voltage insures frequency stability regardless of wide excursions in supply voltage. Although operation of the vfo and final on the same frequency frequently causes chirp, the buffering and isolation features of this design result in a chirp-free vfo. In fact, you can key the entire transmitter and vfo by breaking the B+ leads without any chirp at all! It is wiser, however, to allow the vfo to run continuously while keying the driver and final stages. Since the vfo emits a healthy signal, you will want to turn off the vfo while receiving by flipping the vfo band-switch to the crystal position.

I change bands with a two-pole five-position rotary switch. I use four positions, while the fifth shuts off the oscillator for crystal operation or stand-by. The circuit board has provision for five bands if you wish to use it. You can get bandspread of about 125 kHz by placing a small 33-pF mica capacitor in series with the 30-pF APC vfo tuning capacitor; on 20 and 15 meters, an 18-pF mica is switched into the series circuit for more useful bandspread.

The circuit board is designed for standard size components. The trimmer capacitors are subminiature Elmenco 404 types (since some stores carry only 30- or 40-pF size of this type, these may be substituted if the size of C2 is increased accordingly); the coils are wound on Amidon toroid cores making for extremely compact high-Q tank circuits.



The TX-1 board before any modification or installation.

It can be an interesting task to design and make your own circuit board, as I have done. Besides actually etching a board there are many other construction techniques you can use like no-etch stick-on copper foil patterns, using copper-clad board with insulated stand-offs or insulated islands and the simple technique of using plain epoxy board, drilling holes for component leads and wiring the circuit with short insulated jumpers and the protruding component leads.[1,2,3]

Regardless of the construction method chosen, use care in soldering component leads and use a heat sink when soldering the semiconductors. A no. 60 drill*—the

*Available for 50c from America's Hobby Center, 146 West 22nd Street, New York, New York 10011.

right size for drilling holes for component leads—is invaluable.

To tune and calibrate the vfo, work on one band at a time. After installing the proper L/C combination, temporarily connect the tank to the proper oscillator ports with *short* wires. Find the vfo signal

At that time, cement the toroids to the circuit board. Finally, make an aluminum shield for the vfo and move on to the transmitter module.

transmitter module

The two-stage TX-1 (fig. 2) uses high f_T high-gain devices (MPS6514, 2N4427,

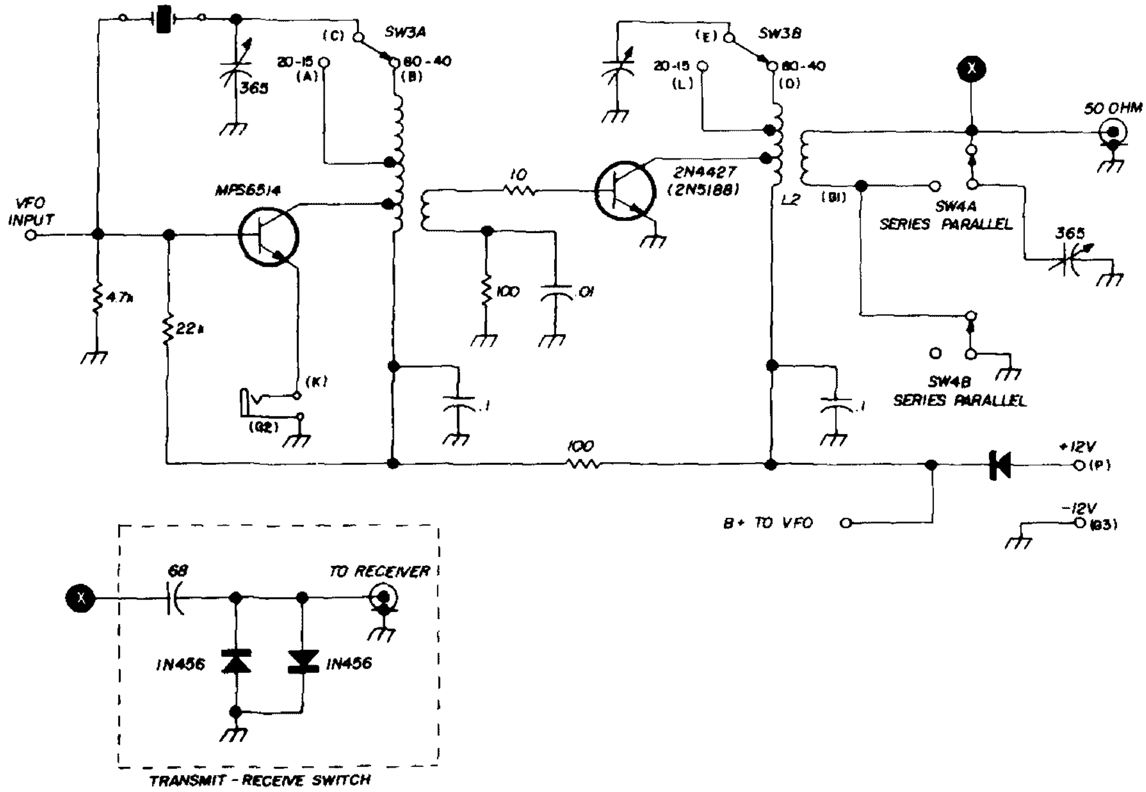


fig. 2. Schematic of the Ten-Tec TX-1 transmitter module. The letters given in parenthesis correspond to the proper port on the TX-1 circuit board.

by listening on the station receiver while adjusting C1. The best way to tune a toroid is to use a variable capacitor, but an adequate compromise tuning method would be to squeeze the turns closer together or spread them further apart. Remember though, this type of adjustment will upset the symmetry of the toroid and can lead to some slight losses and stray radiation. However, some coil tuning will probably be necessary on 20 and 15 meters.

Once you initially calibrate all bands, mount the board, connect the B+, the driver base and the tuning capacitor. Calibrate the vfo for final bandspread *only* after the transmitter is completed.

which can be replaced by the cheaper 2N5188 at 60c), with their collectors tapped down on toroids for proper impedance matching. Preparation of the TX-1 requires two minor modifications. First, remove the trimmer capacitors used in the two tank circuits. To do this, insert a knife blade between the board and one end of the capacitor, melt the solder, and twist the knife blade until the capacitor lead is clear of the board. Second, if you wish crystal control capability, remove the crystal socket *from the board and* mount it on a panel. A third possible modification is the addition of an antenna tuning capacitor across the output link (see schematic). Simply break the

common (ground) copper strip at the bottom of the circuit board before it reaches the ground side port (G1) of the antenna link. Attach the antenna link directly to the dpdt switch, thus allowing you to shift between *series* and *parallel* tuning to accomodate any inductive or capacitive reactances your feedline may present. However, this luxury is not essential and may be omitted without limiting the rig's effectiveness.

To mount the TX-1, simply connect the key leads, B+, vfo output, crystal socket and bandswitch. A word of caution about the bandswitch. Keep the bandswitch leads as short as possible. The capacitance added by four inches of leads, for example, will make it impossible to peak the driver on 15 meters. I recommend mounting the TX-1 flush against the panel where the switch is to be mounted. Two spdt switches instead of a single dpdt will insure the shortest possible leads.

tuning

If you haven't worked with a solid-state transmitter before, a few words of advice are in order. First and foremost, *do not attempt to operate the transmitter without a proper load*! To do so is the easiest way to zap transistors. Further, if your swr is over 3:1, do not hold the key down for long periods – three seconds on is a good rule of thumb. Mismatch can cause thermal runaway – the puncturing of internal transistor junctions because of the heat resulting from the excessive current. Normally the 2N4427 should be comfortably warm. Also watch for self-oscillation, shown by the transmitter continuing to put out rf after the key has been opened. If you encounter this problem, immediately remove the B+. It is usually caused by improper final tuning. Transistors can be tricky devices and they have suicidal tendencies; self-oscillation *inevitably* leads to self-destruction in the transistor world.

Become familiar with the tuning of the TX-1 before making modifications. Install a crystal and jumper wires for the proper band, and tune the driver trimmer until

table 2. Measured outputs from finished rig: (R = 49 ohms)

band	voltage	watts
80	13	1.87
40	12	1.4
20	10	1.02
15	7.5	0.56

Measured using the circuit of figure 3; R = 49 ohms. P = V2/2R

the crystal breaks into oscillation. Peak tuning is accompanied by pulling of the crystal frequency. With a 50-ohm non-inductive load and rf indicator (fig. 3) connected across the output posts, close the key and tune for an output indication. Retuning the driver and final will be necessary for peak output. Remember the 2N4427 is not a 4-1000; it should be

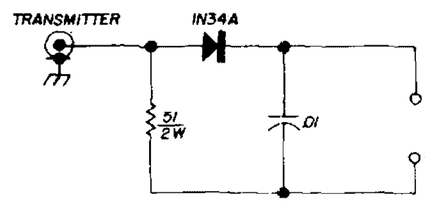


fig. 3. Simple rf output indicator for use with almost any QRP transmitter. Voltmeter should read at least 15 volts full scale for use with this rig.

respected and not pushed unnecessarily. Tuning with the vfo is about the same as with a crystal. The final tuning with the vfo will result in pulling the vfo frequency several hundred hertz when the key is closed. This is normal, and peak output should occur with a pulling of about 500 hertz.

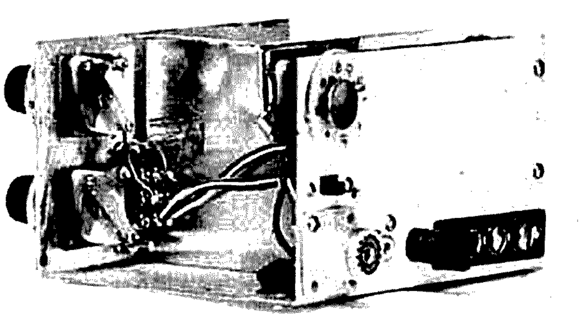
construction

The photos show my approach to construction. I housed the transmitter in a 3 x 4 x 5-inch box, making for some tight fits. The only critical aspect of mounting is the drive/final bandswitch. I used cheap mica-dielectric variable capacitors, and they work well. The switch used to insert series capacitance to the vfo tuning capacitor should be a good toggle

switch — slide switches usually exhibit intermittent contact, a detriment to vfo stability. A small vernier and 30-pF APC capacitor provide smooth tuning on all bands. Since all critical parts are mounted on the circuit boards, you can work out your own mounting setup.

operation

It is hard to define what is so fascinating about QRPP. I suspect it is accepting the challenges of QRM and propagation, and with a dry cell and a transmitter that



Rear view of the completed transmitter shows use of inexpensive bc type capacitors and the shield around the vfo compartment.

fits into a lunchbox, working the world. In a sense, it is getting back to the basics that characterized most of our Novice experiences. The story told by converted appliance operators is always the same: the KWS-1 gathers dust while the QRPP rig is worked to death providing all the excitement that ham radio can offer. And the thrill never seems to leave — you can work a KH6 on 40 meters three nights in a row (as I did this spring) with a couple hundred milliwatts, and you still tremble as he comes back to your call, you still feel the exhilaration of man stripped to the bare essentials confronting and overcoming nature. It's a great experience!

Let me try to give you a realistic idea of what to expect from QRPP operation and how to go about it. QRPP operation requires skill, patience and an understanding of how everything in the transmitting system works together. Propagation is important — what to expect of your rig and antenna in terms of distance and signal strength during a given season at a particular time of day on a particular band. Likewise, the QRPP operator attempts to be as efficient as possible in his overall system — good matching, efficient feedline, accurately cut antenna, clean signal. In short, he attempts to offset his power disadvantage by knowledge, skill, and above all, patience.

Every newcomer to QRPP is amazed at what can be done with a few hundred milliwatts; for example, newcomer WA8WWS, during the first ten days of operation at 2.5-watts input, managed to work 54 stations in 18 states on 40 meters. Some guys turn into fanatics upon seeing what QRPP can do — WA8DDI was so excited that after six months he has WAS and over 50 countries with his 1-watt output rig! But for most of us, QRPP is just the way we approach ham radio for daily enjoyment of the hobby — ragchews, casual contacts and the like. Here are some brief suggestions to help you master QRPP operating skills.

You can assume that most stations will be running high power, and if a station calling CQ is weak, it is very likely that propagation between his location and yours is bad. Instead of calling him, find a stronger signal that you can copy well. Experience has shown that calling CQ with QRPP is futile — although, it doesn't hurt to try as long as you don't expect a logjam of answers! Third, select your transmitting frequency carefully. Many hams use transceivers with ±1-kHz offset tuning, so this is normally the limit of deviation for a calling station. But even in a crowded situation, this is usually enough to put your signal into the open. If interference appears after making contact, suggest changing frequency, but let him call you so you can zero beat him. Cultivate a good clean fist and send at medium speed until asked to speed up — it is easier to copy a slow signal than a fast one. Don't be afraid to repeat things two or three times if the going is

rough — the other guy will appreciate it. Inform the other station that you are QRPP immediately upon making contact — this inevitably causes him to turn up the rf gain and notch the filter a bit deeper. Most amateurs find it exciting to work a milliwatt station; take advantage of it. Similarly, always afix QRP to your beginning and concluding signature, at the end of a contact, give a short QRZ — it's amazing how many fellows will read the mail when they hear someone sign QRP, and will gladly come back to a QRZ to give you a report. Don't waste all night calling the same station — if he hasn't come back on the third call, he probably has the rf gain turned back waiting for a blockbuster to shake his shack. Don't be shy — usually a station calling CQ DX will gladly accommodate a milliwatt station, and the psychology is perfect — he's tuning for weak signals in the first place, and that's you! Above all, be *patient* and know what to expect from your gear and propagation phenomenon. You will find that after getting the hang of it, it will be possible to do almost as much with QRPP as with high power. In populous areas of the country, QRPP stations have as high as an 85% call-to-contact ratio. The main point is the QRPP operator voluntarily imposes upon himself a power limitation, but after getting the hang of operating QRPP, wonders why he ever thought he needed the kW amplifier in the first place.

Whether you are a newcomer or oldtimer, this is a good rig for the investment of time and money, and it has performed flawlessly for me. Give yourself a break — try QRPP.

reference

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ham radio

using Y parameters in rf amplifier design

By using Y parameters in rf circuit design, the designer can determine stability and gain before building a breadboard

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On the data sheet for the RCA 40673, a dual-gate mosfet, is the statement, "The reduced capacitance allows operation at maximum gain *without neutralization."* This is a comforting thought when looking for a suitable rf amplifier. But then you stumble onto a circuit in the 1970 ARRL *Radio Amateur's Handbook* for a dual-gate mosfet rf amplifier, and it is pointed out that neutralization is usually required.

The truth is, both assertions may be correct. The secret of the stability or instability of a device is locked up in its *y parameters;* the purpose of this article is to give some explanation of them and practical instruction in their use. Let's define some basic notions.

The quantities resistance, r, reactance, x, and impedance, z, should be generally familiar. If not, there is a good primer in the ARRL *Handbook.* The reciprocals of these quantities are very useful and have the following nomenclature and symbology:

$$\text{Conductance} = g = 1/r \quad (1)$$

$$\text{Susceptance} = b = 1/x \quad (2)$$

$$\text{Admittance} = y = 1/z \quad (3)$$

Note that these quantities are measured in mhos, derived from ohm spelled backwards!

Understanding the relationships between these six quantities is not too difficult, however. The standard way of handling them mathematically is to use the complex plane and complex algebra. It's not too much fun on a slide rule, but quite necessary to get the answers. Since the y parameters are all complex, and the stability calculations involving them must be carried out using complex arithmetic, I have added an appendix on complex numbers with an example or two. Don't let the word "complex" scare you. High-school math books should have additional information on them.

y parameters

In this discussion, I will not deal with all the whys and wherefores that design engineers use in the rf amplifier problem. Suffice it to say, that although impedance *could be directly used in design, and* although other parameters (like the s and h parameters) could be used, the y parameters make the work the easiest, and are commonly found on device spec sheets. Let's begin by defining them.

Assume for discussion that the transistor is a "black box" sketched in **fig. 1** to which is connected an rf signal source and an output load. *In a real circuit you* must apply power and bias, etc., but these are ignored for this simple analysis. Likewise, you might, in real life, have an antenna for the signal source, or generator, and an output tuned circuit coupled to a mixer for the load. But let's be completely *general* and simply specify that at the input of the black box there is a generator of voltage e_1, which *sends a* current i_1 into the box.

Seemingly a bit strange at first, the load is also viewed as a *generator.* Why not? A voltage e_2 appears across it with its associated current i_2. To a design engineer, the "generator" of voltage e_2 producing a current i_2 is a perfectly logical analysis tool. You may get a bit more feeling for this view if you remember that coupling energy from the load to the input (the load now becomes a signal

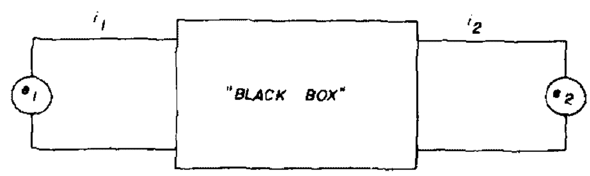

fig. 1. Y parameters of a transistor are defined in terms of the input and output voltages and currents.

source or *generator)* is *precisely* what is needed to make an *oscillator.* The truth is, there *are* coupling paths through the transistor *itself – completely divorced* from your external circuit layout – which will make your rf amplifier *an oscillator, and that's precisely what* this article is trying to help you avoid!

Let us now simply state some results which can be derived mathematically from ac circuit theory, namely

$$i_1 = P_{11}e_1 + P_{12}e_2, \text{ and} \quad (4)$$

$$i_2 = P_{21}e_1 + P_{22}e_2 \quad (5)$$

The mathematical coefficients or "P" parameters in these equations are characteristic of what's in the black box (i. e., the transistor) and are constant except for changes in bias, temperature or signal frequency.

Let's explore the first equation a little. Suppose we *short circuit* the output. The voltage e_2 must then be zero. (Would you hear that VU2 with a jumper across your rf stage output?) The equation then reduces to $i_1 = P_{11}e_1$ under these circumstances. We could also write this as $e_1 = i_1(1/P_{11})$ which is nothing more than Ohm's law in slight disguise. What usually goes in place of the $(1/P_{11})$ is

be measured in *mhos,* and it is one of four similar parameters (including P_{12}, P_{21} and P_{22}) in the equations.

Since the parameters are all in mhos – that is units of admittance – they are called admittance parameters or y parameters. Having thus identified them the equations may be written:

$$i_1 = y_{11}e_1 + y_{12}e_2 \quad (6)$$

$$i_2 = y_{21}e_1 + y_{22}e_2 \quad (7)$$

When a spec sheet gives you a number value for a y parameter, it must also specify the operating point, frequency, etc. Curves of the y parameters against frequency or bias are often given.

How does the transistor manufacturer get values for the y parameters published on the spec sheets? Several manufacturers make admittance bridges which do the job and are capable of measuring over a wide range of frequencies – *even into the* GHz region. Some use capacitors for output ac short circuit, others use tuned lines. Neither short upsets the dc operating point, and, of course, a large number of devices must be measured to get the so-called typical values.

Why should the y parameter be measured by shorting the *device? Mathe*matically, either e must be eliminated by short circuit, or *i* by open circuit to make the equations solvable. At radio frequencies a *true rf* short-circuit is easier to make than a true open-circuit, so making measurements becomes more practical this way. As hinted before, there are other parameters such as z parameters for impedance, h for hybrid, etc. However, if you have a complete set of one kind of parameter, it can be mathematically transformed into any other kind.

Therefore, the definitions for the y parameters can be written using **eqs. 6** and **7**

$$y_{11} = \frac{i_1}{e_1} \text{ when } e_2 = 0 \text{ (output shorted)} \quad (8)$$

$$y_{12} = \frac{i_1}{e_2} \text{ when } e_1 = 0 \text{ (input shorted)} \quad (9)$$

$$y_{21} = \frac{i_2}{e_1} \text{ when } e_2 = 0 \quad (10)$$

$$y_{22} = \frac{i_2}{e_2} \text{ when } e_1 = 0 \quad (11)$$

These equations apply to any "linear active two-port (that is, input and output) network," (LAN) and are good for bipolar transistors as well as fets and IC rf *amplifiers. However, when applied to* fets, the number subscripts have yielded to descriptive letters which I will use since the example design will use an fet. **Table 1** cross references and names the parameters. The s in all the designations refers to a common-source configuration.

Note that the y parameters are complex quantities, that is, $y_{is} = g_{is} + b_{is}$ or, input admittance is the sum of input conductance and input susceptance.

Let's see what we can make of these. Looking into the fet amplifier (i.e. gate) you would see both resistance and capacitance. The same would hold true for a circuit looking into the fet output (i. e. drain). These impedances (resistances and capacitive reactances) can be expressed as admittance according to **eq. 3** and are exactly what y_{is} and y_{rs} are when appropriate short circuits are made as outlined before.

Y_{fs} is similar except that it relates to *output current* and *input voltage* and is therefore transadmittance. Remember transconductance in vacuum tubes? Y_{fs} is essentially the same but includes the reactive part too.

Y_{rs} involves output voltage and input current and is the path for signals back through the device. Remember plate-grid capacitance and neutralization of your rf amplifier? Y_{rs} is much the same but includes the non-reactive (i. e. conductive) part of the feedback path as well.

You see, a properly designed vacuum tube normally has no conductive path for

electrons from plate to grid, only capacitance. But a transistor does have a conductive path back through the semiconductor material, although it is negligible for many purposes. Y_{rs} then includes both the resistance and capacitance, expressed in terms of conductance and susceptance.

stability

Several years back, J. G. Linvill[1] devised a method of determining the stability of a device using the y parameters. The Linvill stability factor C is given as

$$C = \frac{|y_{12}y_{21}|}{2g_{11}g_{22} - Re(y_{12}y_{21})} \quad (12)$$

Absolute values and real parts are discussed in the appendix (g_{11} is the real part of y_{11}, etc.). When C is less than 1, the device is unconditionally stable. If it is greater than 1, it is potentially unstable.

Since the Linvill stability is taken for the worst possible case, that of infinitely large source and load resistances (i. e. open circuit), there arises the possibility of rendering the potentially unstable device tractable. Note that I say *device.* This stability refers to signal paths through the transistor, not to paths due to stray circuit capacitance, etc. A. P. Stern[2] has defined the Stern stability factor K which includes the effects of input and output loads as

$$K = \frac{2(g_{11} + G_S)(g_{22} + G_L)}{|y_{12}y_{21}| + Re(y_{12}y_{21})} \quad (13)$$

G_S and G_L are the conductances of the source and load impedance respectively. If the value of K is less than 1, the amplifier is unstable. Values of K around 2 to 4 should be satisfactory for a well laid out amplifier to be stable. For some devices at some loads K values over 100 appear.

The first step in amplifier design then, is to compute C. If it is less than 1, stable

table 1. Y parameters for field-effect transistors. The subscript s indicates a common-source configuration.

$y_{11} = y_{is}$	input admittance
$y_{12} = y_{rs}$	reverse transadmittance
$y_{21} = y_{fs}$	forward transadmittance
$y_{22} = y_{os}$	output admittance

design in easy, and only external feedback paths must be eliminated by proper layout. If C is greater than one, you must compute K, adding in the source and load conductances which are frequently derived from the resonant impedance of a parallel tuned circuit.

The tuned circuit is designed for inductance, capacitance and loaded Q values compatible with the selected goals in impedance matching, bandwidth, etc. The tuned circuit design may lead you into conflict, however. Maximum gain occurs when the source and load are matched to the transistor. However, for many purposes quite wide mismatching affects the gain by few enough dB that gain can be sacrificed to achieve other objectives.

Indeed, deliberate mismatching is one way to achieve a large K and a stable amplifier. You can see in **eq. 13** that as the source and load resistances decrease, G_S and G_L increase, and, being in the numerator, increase K. So, smaller source and load impedances make for a more stable amplifier.

To see how badly mismatching affects gain, you can compute the gain from the y parameters:

$$\text{Gain} = \frac{|y_{21}|^2 G_L}{|Y_L + y_{22}|^2 \, Re(y_{11} - (y_{12}y_{21}/(y_{22} + Y_L))} \quad (14)$$

For a tuned circuit at resonance Y_L equals G_L.

feedback

A second method may be used to achieve stability – *feedback.* There are two kinds provided by proper feedback networks. The first is unilateralization

which reduces y_{12} to zero. Since there is then no input/output communication, the amplifier is stable if well laid out, and a bonus is that tuning the output can't detune the input (good input/output isolation).

The second feedback scheme is neutralization which reduces y_{12} to some value other than zero. The common circuit with a neutralization capacitor wipes out the reverse transsusceptance, but not the reverse transconductance, so y_{12} is not zero. Maximum input/output isolation is not achieved, but perfectly sufficient stability may be.

In summary then, mismatching is the easiest way to achieve stability since it requires no additional circuitry, but is achieved at the expense of gain. This loss may be small, and quite tolerable, however. You have seen reference to detuning to make an rf amplifier stable; this is pretty crude mismatching. It is likely to be quite costly in gain, and is no substitute for selection of optimum resonant circuit impedances which will meet the various stability, bandwidth and other criteria.

Finally, before some examples, there are several other useful expressions which involve y parameters and are used by the sophisticated design engineer. I will mention only two in passing, the input and output admittance equations:

$$Y_{in} = y_{11} - \frac{y_{12}y_{21}}{y_{22}+Y_L} \tag{15}$$

$$Y_{out} = y_{22} - \frac{y_{12}y_{21}}{y_{11}+Y_S} \tag{16}$$

Normally, in a tuned rf amplifier, the reactive (imaginary) parts of Y_{in} and Y_{out} simply become part of the tuned circuit reactances and disappear when tuning for peak response. The real parts G_{in} and G_{out} load the tuned circuits and enter into their design for proper bandwidth; that is, they contribute to the value of loaded Q. They are expressed as equivalent resistances which appear in parallel with the tuned circuit and figure into the calculations for circuit Q.

practical design

I will now give a design example with some numerical values with which you may check your understanding of complex calculations. A set of y parameter curves is given for the MPF121 dual-gate mosfet in **fig. 2**. The values are in millimhos; the answers will be also. Let's try the design of converters for 2 and 6 meters with a 10- to 14-MHz tunable i-f provided by the station receiver. Reading off the values at 50 and 150 MHz, for 6 meters we get:

$$y_{is} = 0.08 + j1.2$$
$$y_{rs} = 0 - j.0065$$
$$y_{fs} = 12.9 - j1.4$$
$$y_{os} = .08 + j.7$$

and for 2 meters:

$$y_{is} = .69 + j4.5$$
$$y_{rs} = 0 - j.023$$
$$y_{fs} = 12.1 - j4.7$$
$$y_{os} = .28 + j2.1$$

First compute the Linvill stability for these two cases. It turns out to be 0.60 for 2 meters and 3.9 at 6 meters. Thus, the MPF121 is unconditionally stable at 144 MHz and conditionally unstable at 50 MHz. Let's proceed with the simpler design first, for 2 meters.

two-meter design

Using normalized tuned-circuit response curves *such as those in the Radio Amateur's Handbook,* and starting somewhat arbitrarily with a reactance of 150 ohms, you find that the required inductance is 0.16 μH, with a capacitance of 7.3 pF. For a response across the 144-148 MHz band within 1 dB down, the required loaded Q is 16, and the parallel resonant impedance of the tuned circuit is 2400 ohms.

This last fact means that the total effective parallel resistance across the tank must be 2400 ohms if our criterion of 1 dB response is to hold. This resistance includes that contributed by the circuit components, that transformed from the antenna, and that due to the input resistance of the fet. The latter may be computed from the y parameters according to the formula for Y_{in}.

Y_{in} of course consists of a real part, the input conductance, and an *imaginary* part, the input susceptance. Taking the reciprocal of the conductance gives the input resistance, while the reciprocal of the susceptance yields the input capacitive reactance. (Remember that millimhos yield answers in kilohms.) The susceptance may be converted to a capacitance using $C = 1/(2\pi f X_c)$, and turns out to be 4.9 pF. This is part of the 7.3-pF tank capacitance, a goodly part indeed; you may need to recalculate for a higher C circuit.

The input resistance is quite low at this frequency, only 1200 ohms. Suddenly you realize that this will degrade the Q if the fet is connected across the whole tank. This may not hurt; however,

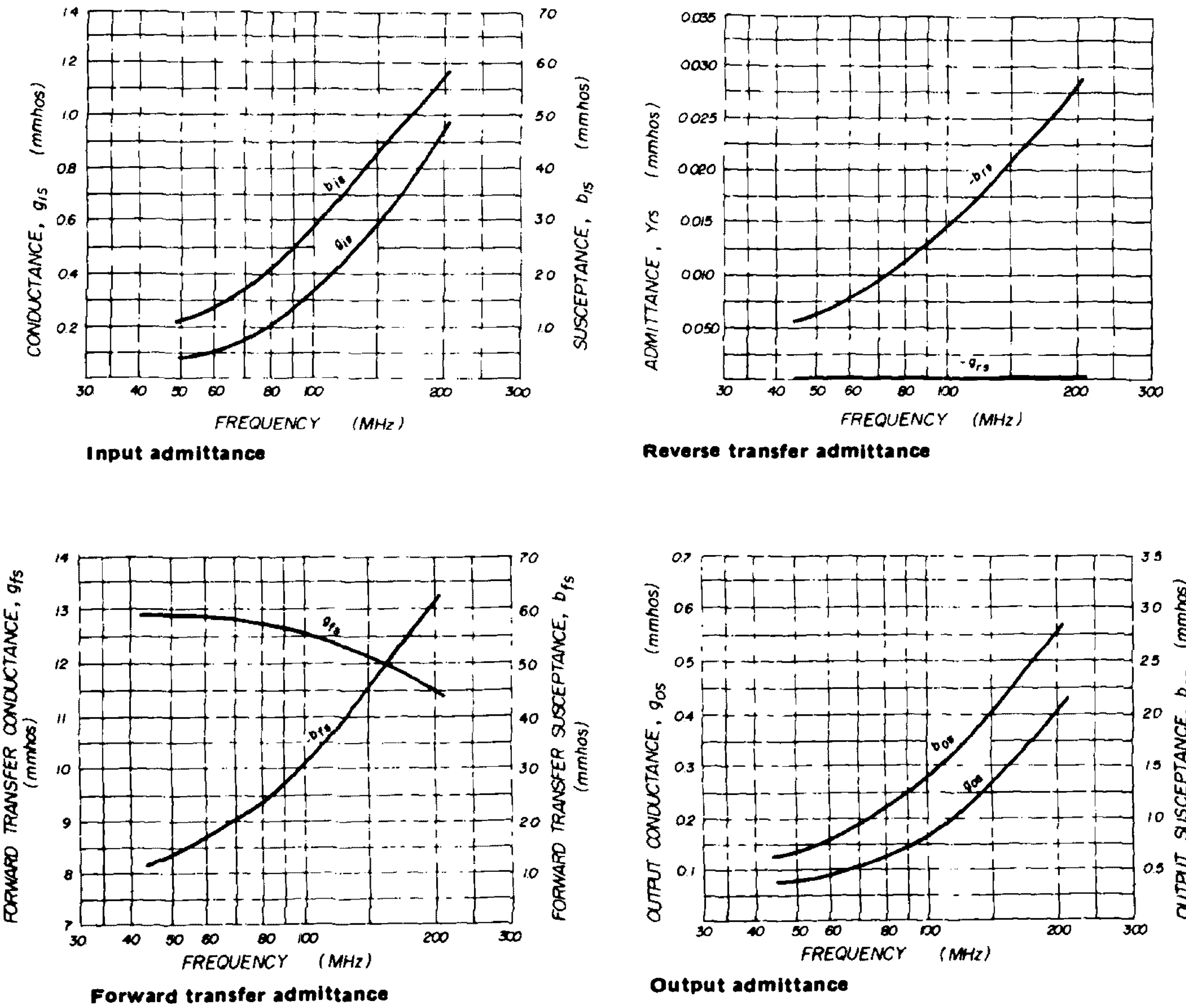

fig. 2. MPF121 dual-gate mosfet common-source admittance parameters (V_{DS} = 15 Vdc, V_{G2S} = 4.0 Vdc, I_D = 6.0 mA dc).

you should look for possible poor image response.

At Q = 16, the response curves show that an image at 122 MHz would be 15 dB down. Using a second identical tuned circuit at the output would yield a ± 1 dB response across the band and image response of -30 dB. You would most likely want to tap the fet input down on the tank to preserve the Q under these circumstances.

Similar procedures apply to the output circuit with fet output resistance of 2900 ohms and output capacitance of 2.3 pF. *Incidentally*, the Stern stability for 2400 ohms source and load resistances turns out to be 8.1 and the calculated gain 12.4 dB. The fet is thus very stable, but it is up to the builder to avoid external feedback paths in his layout.

six-meter design

Let's look at the 6-meter case next. Starting with a reactance of 150 ohms, and ± ½ dB response across 50 to 51 MHz (i. e., 1 dB down at the band edges), you find that L = 0.47 μH, C = 21.0 pF and loaded Q must be 22.2. This gives a resonant impedance of 3333 ohms, and an image response at 30 MHz of -25 dB for the single circuit. If you use identical input and output tanks, the response across the band segment would be ± 2 dB and the Stern stability 3.8, which is adequate. The calculated gain is 26.5 dB. Suppose the slug-tuned coils had unloaded Qs of 80. R = QX = 80(150) = 12k which is the resistance contribution of the coil alone. The input resistance of the MPF121 is 5620 ohms, so the following parallel resistance relationship must hold:

$$\frac{1}{5620} + \frac{1}{12K} + \frac{1}{R_{ant}} = \frac{1}{3333} \qquad (17)$$

Solving, R_{ant} must be 26k. Thus, the antenna (perhaps 50 ohms) must be transformed to 26k by the input circuit in order to preserve the loaded Q of 22.2. Frankly, this makes an antenna coupling of rather poor efficiency.

The design has several alternatives at this point, some of limited usefulness, including: tap the fet down on the tank; use a higher Q coil; accept a lower Q through tighter antenna coupling with its wider bandwidth and poorer image rejection; use a coupled circuit to gain selectivity and largely separate the fet and antenna loadings. Choices like these must be weighed to arrive at a final good design.

Suppose you wish to use a transistor for a high-frequency receiver, but no y values are given below 30 MHz? After inspecting a number of data sheets, a few of which had values for lower frequencies, and assuming that dual-gate mosfets have *similar* general characteristics, even though they are advertised for a given frequency range, I conclude that the following rules-of-thumb would be better than using single values for the high-frequency region:

1. g_{rs}, g_{fs} and g_{os} are constant throughout the hf region.

2. The four susceptances may be approximated assuming they are proportional to frequency. That is, a b_{is} of 1.1 at 40 MHz would be close to 0.4 at 14 MHz. Use the parameter value at the lowest frequency given on the spec sheet as a starting point.

3. g_{is} appears to fall off by a factor of four per octave. For instance, a value of 0.24 for 10 meters would reduce to 0.06 for 20 meters, 0.015 for 40 meters, etc.

Note that y_{is} means *input* admittance in the common-source configuration. What about common-gate and common drain? Formulas to convert the common-source parameters to the other configurations, Y_{ig}, Y_{id}, etc., have been worked out and are given in reference 4. All equations for stability, gain, etc. are applicable to the common-gate and common-drain configurations as they stand, simply by using the appropriate y parameters for the configuration of interest.

summary

I may summarize then, with some

rules-of-thumb which may prove beneficial even if you make no calculations:

1. A typical dual-gate mosfet will be unconditionally stable if the frequency is high enough.

2. A potentially unstable device may be stabilized by feedback or mismatching.

3. Mismatching may be preferable for stabilization since it requires no extra components, no adjustment, is not frequency sensitive, and the cost in gain is usually tolerably small for amateur work.

4. The smaller the source and load impedances, the greater will be the stability. Thus, at parallel resonance, smaller values of both reactance and Q give smaller parallel impedance and greater stability.

In conclusion, I must point out that there are techniques in rf amplifier design used by the professional design engineer which have not been mentioned here. Some of these methods require access to a computer since the number of computations may be vast. In addition, I have not pursued the design examples through to final optimization, but rather pointed out procedures and design choices. As you can easily see, there are tradeoffs between many factors such as selectivity, high gain, optimum noise figure and reasonable cost which a designer must evaluate.

Ultimately, the stability criteria are of key importance since no one can accept an unstable oscillating rf amplifier! Although I was privileged to program and run my calculations on a computer, I hope that someone may be brave enough to tackle the calculations and possibly avoid generating his own input signals in that new receiving system!

appendix

complex numbers

The square roots of negative numbers are designated *imaginary numbers* and there are definite algebraic rules for dealing with them. The square root of minus one is the basis for complex operations, and it is generally given the symbol i in mathematics while engineers use j. Thus, $i = j = \sqrt{-1}$, and from the rules governing square roots, you can verify that $j^2 = -1$, $j^3 = \sqrt{-1}$, $j^4 = 1$ and so forth. Ordinary algebraic rules also apply such as $(j)(j) = j^2$, $j^3/j = j^2$, etc. Any nonimaginary number is a *real* number.

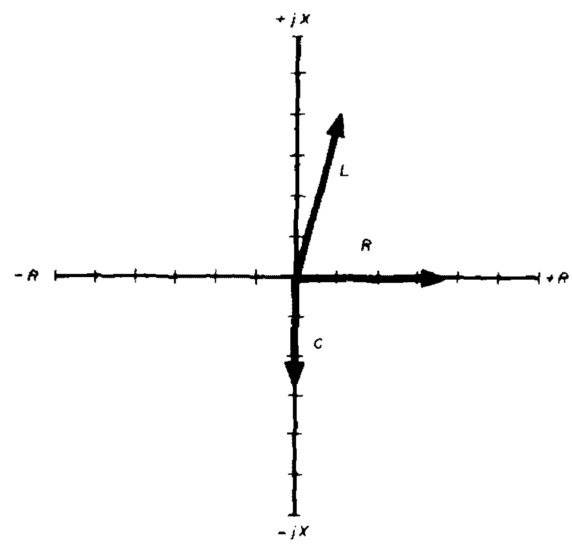

fig. 3. Circuit elements L, R and C plotted on a complex coordinate system. Each division is 100 ohms.

A *complex number* is the sum of two numbers, the first real and the second imaginary, such as $3 + j2$, $2.5 - j7$, or in general, $a + jb$. The properties of j and complex numbers make them suitable for describing the characteristics of resistors, capacitors and inductors plotted on coordinate axes on the *complex plane*. These plots make it easy to visualize the behavior of these circuit elements, and the associated complex algebra makes it possible to determine their behavior with slide rule or computer.

The basic reason that the complex plane is needed is that the voltage is 90° out of phase with the current in any pure reactive element. Thus, in **fig. 3**, the complex coordinate system is shown with three circuit elements plotted thereon, all

at a frequency of 60 Hz. A pure 350 ohm resistor is shown at R. It lies on the R axis and has no component along the jX axis. A pure 10 μF capacitance is labeled C and since $X_C = 1/(2\pi fC)$, it has 270 ohms capacitive reactance at 60 Hz. The voltage across the capacitor is 90° behind the current, and the capacitive reactance is along the negative jX axis.

The 1.06-henry inductor is not pure since the wire has a resistance of 100 ohms. (Any real capacitor has *some* resistance associated with it too!) The inductive reactance is $+jX = +j(2\pi fL)$ (positive since the applied voltage leads the current by 90°) and has a value of 400 ohms. The inductor then has two components, 400 ohms of pure inductive reactance and 100 ohms of pure resistance (plus some stray capacitance in any real inductor which we ignore). These two quantities have the properties of vectors, and are shown plotted in fig. 4.

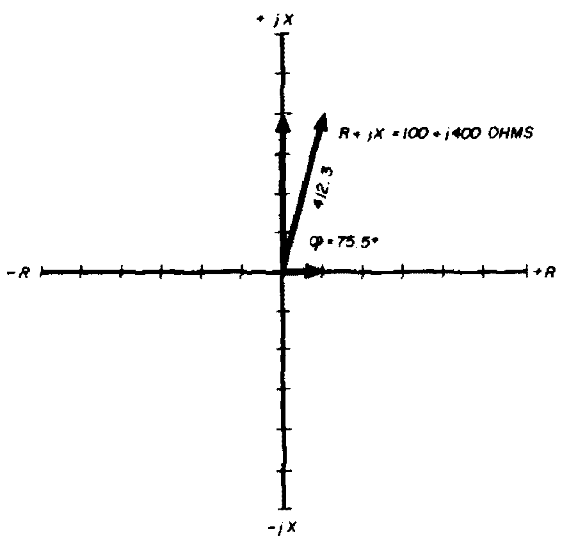

fig. 4. R + jX characteristics of 1.06-H inductor discussed in text. Each division is 100 ohms.

The two components R and jX lie along the axes while the representation of the inductor $R + jX = 100 + j400$ ohms is the resultant of the two, and is a complex number. We say that the device has 100 ohms resistance plus 400 ohms inductive reactance. The sum of these (and it is a *vector* sum) has a magnitude obtainable by the Pythagorean theorem as follows:

$$\sqrt{100^2 + 400^2} = \sqrt{170{,}000} = 412.3 \text{ ohms}$$

It also has a phase angle θ which is 75.5°.

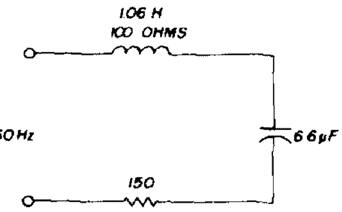

fig. 5. Simple RLC circuit. Complex characteristics of this circuit are plotted in fig. 6.

Both the complex representation $100 + j400$ ohms and the vector form $412.3 + \angle 75.5^\circ$ ohms are equivalent and are simply different ways of stating the *impedance* of the inductor Z_L. We may then state the general relationship that $Z = R + jX$.

As an example of how the behavior of a circuit can be visualized, plot the characteristics of the circuit in fig. 5 on the complex plane. The capacitor has a reactance of $-j400$ ohms. All components are shown in fig. 6A. All components are added vectorially together to find the result; you can see that $-j400$ cancels $+j400$ while $150 + 100 = 250$ ohms. This can be done by computation alone as follows:

$$Z_{total} = Z_R + Z_C + Z_L$$
$$= 150 - j400 + 100 + j400$$

Real and imaginary parts are grouped together giving

$$Z_{total} = (150 + 100) + (j400 - j400) = 250 \text{ Ohms}$$

If you change the frequency to 120 Hz,

$$Z_{total} = Z_R + Z_C + Z_L$$
$$= 150 - j200 + 100 + j800$$
$$= 250 + j600 \text{ ohms}$$

and at 30 Hz

$$Z_{total} = 250 - j600 \text{ ohms}$$

These values of L and C are series resonant at 60 Hz since the inductive and capacitive reactance cancel at series resonance; the current is determined only by the resistor and the resistance in the inductor. Incidentally, you may relate the negative R axis to *amplification.*

In the preceeding manipulation of complex numbers, no special techniques were required. But multiplication and division must be added to the list. Let's try an example by multiplying the following complex numbers:

$$(2 + j3)(3 - j5)$$

This operation is carried out just as the algebraic multiplication of two binominals:

$(a+b)(c+d) = ac+ad+bc+bd$
$(2)(3) = 6$
$(2)(-j5) = -j10$
$(+j3)(3) = +j9$
$(+j3)(-j5) = (+j)(-j)(+3)(+5) = -(-1)(15)$

Adding all terms, we get $(21 - j1)$

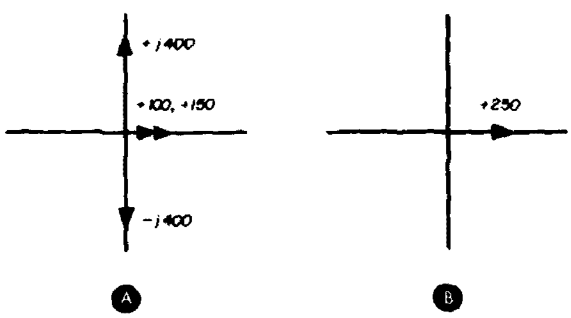

fig. 6. Complex characteristics of the circuit in fig. 5.

Division of complex numbers requires the use of the *complex conjugate* to get rid of the imaginary parts in the denominator. For any complex number $(a + jb)$ its complex conjugate is defined as $(a - jb)$, and vice versa. Let's see what happens when a number is multiplied by its conjugate, remembering that $(a + b)(a - b) = a^2 - b^2$:

$$(a + jb)(a - jb) = a^2 - (jb)^2$$
$$= a^2 + b^2$$

The j's drop out. Division is performed by multiplying the denominator by its complex conjugate, and also multiplying the numerator by the same quantity so that the value of the fraction remains unchanged as follows:

$$\frac{4 + j8}{2 + j6} = \frac{(4 + j8)(2 - j6)}{(2 + j6)(2 - j6)}$$
$$= \frac{8 - j24 + j16 + 48}{4 + 36}$$
$$= \frac{56 - j8}{40} = 1.4 - j0.2$$

You thus get rid of j's in the denominator and can easily get a simple answer.

Two additional complex manipulations will be necessary. The first is commonly designated $Re(a + jb)$, and means the real part of $(a + jb)$, or just a. Thus $Re(5 - j2) = 5$. Secondly, $a + jb$ designates the absolute value. It is always positive, and by definition, is the magnitude of the quantity when expressed in vector form. Thus

$$|5 - j2| = \sqrt{5^2 + 2^2} = \sqrt{29} = 5.4$$

For the inductor in fig. 4, $(100 + j400)$ ohms, you may write $Re(100 + j400) = 100$ and $|100 + j400| = 412.3$

In summary, addition, subtraction, multiplication, division and the evaluation of real and absolute values just about covers the needed techniques to get started with complex numbers in a useful way.

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ham radio

ham radio sweepstakes winners

W3PTG wins grand prize, puts all new Drake station on the air; WAØKKC proudly using new Standard fm station on two meters

Skip Tenney, W1NLB, Publisher, Ham Radio Magazine

Ham Radio Magazine's 1972 Sweepstakes is now history, and it was a busy bit of history at that. After many months of hard work by our staff here in Greenville (have you ever tried to open and read over a thousand pieces of mail in one day?), the big drawing finally took place as scheduled on May 18th. As you can see, we had quite a box full of entries. In fact, it was a large job just to mix them up thoroughly so that everyone had an even chance.

Many of the people who bring you HAM RADIO each month gathered to watch Pat Hawes draw the name of this year's Sweepstakes winner—Gus Haak, W3PTG.

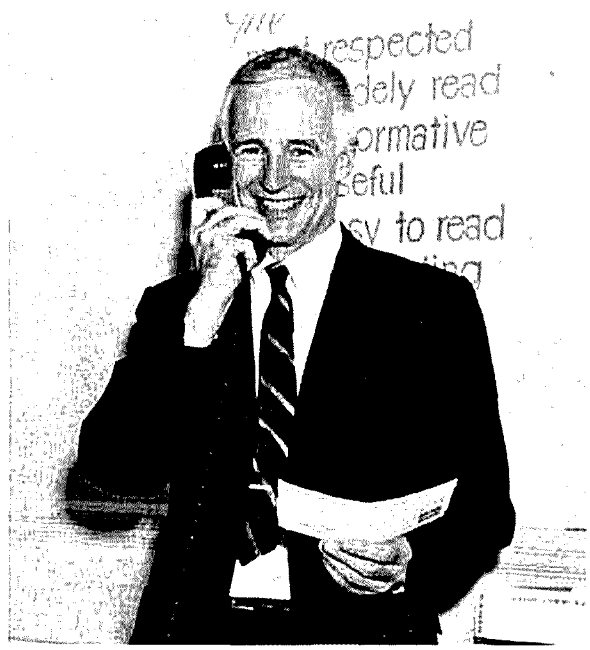
winners of Radio Communications Handbook

W1VLD	W5BL	W9IQN
WA1FXA	W5RDE	W9JQY
WA1MFI	K5SZH	W9JYY
W2PKY	WA5OQR	K9ALD
W2VLS	W6CTY	K9GET
WB2QKQ	K6CMV	K9RYW
W3BSE	K6SEQ	WA9GAY
W3GWM/1	WA6PQG	WA9SCD
W3MCL	WA6TMQ	WB9ADL
W3TOL	W7MUG	WN9FIJ
W3UZN	WA7BKW	WØCOE
WA3PJL	W8WRJ	WØZLO
W4KQK	W8YBM	KØKKK
K4EHP	K8ZPR	KØOJR
K4SHJ	WA8ODV	WAØNUG
WB4VGR	WB8BOT	WAØUEN
WN4ZCF	WB8LAT/9	

The winner of the grand prize was Gus Haak, W3PTG, who by now has had his new Drake TR-4, RV-4 and L4-B in operation for over a month. Gus was a perfect choice for this gear as he is active on 10, 15 and 20 meter sideband and was running a considerably more modest station before now.

This new station provides him with an outstanding signal on all the hf bands. The Drake TR-4 is one of the most respected ssb transceivers on the market while the RV-4 allows split frequency operation. The L4-B linear permits 2 kW PEP ssb operation and 1 kW a-m, CW and RTTY on a continuous-duty basis.

Our second prize went to Dick Mollentine, WAØKKC. Again, the prize was ideally suited to it's winner's needs. Dick has primarily been active on 6-meter a-m until now, and he had been eyeing 2-meter fm with great interest.



Ham Radio publisher Skip Tenney, W1NLB, telephoning W3PTG to tell him he has just won the 1972 GRAND PRIZE.

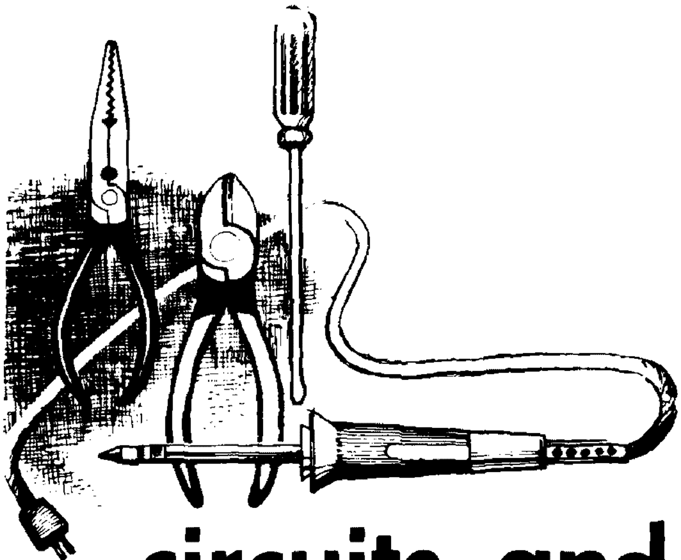
Assistant editor Doug Stivison, WA1KWJ, enjoys a contact before sending off Standard's versatile SR-C146 handheld fm transceiver to this year's winner.

With his Standard trio consisting of a SR-C146 hand-held 2-meter fm transceiver, SR-C826M mobile rig and the all-new SR-C14 base station, he can claim to have the perfect piece of fm gear for every operating requirement. Of particular interest is the base station. When the prize was awarded it was one of the first of this model to arrive in this country!

The 50 third prizes of Radio Communications Handbooks pretty well covered the whole country as you can see from the list of winners. We were glad to find that the list included three YLs and two novices, along with a wide variety of calls, both old and new.

Again, we would like to thank the many thousands of you who entered. Someday we hope to figure out a way so that everyone can win.

ham radio



circuits and techniques

ed noll, W3FQJ

ic flip-flops

Last month this column contained a series of experiments that permit you to observe the operational characteristics of various gates and a basic flip-flop multivibrator. Data on the construction of a 100-kHz calibrator oscillator and the use of a universal flip-flop as a two-to-one counter was presented. In this month's column, specific experiments permit you to take a more detailed look at the operational characteristics of a universal (JK, R-S and clocked) digital multivibrator.

universal flip-flop logic

The very simple circuit arrangement of **fig. 1** is appropriate for checking the logic voltages at the various terminals of the SN7472 J-K master-slave flip-flop. Use Fahnestock clips or binding posts as a convenience in making circuit changes. The unit has three sets of J and K inputs; only one set need be used. Other inputs are R, S (clear and set) and clock. At each of these inputs, logic 1 is the no-connection condition. Logic 0 is established by shorting any one of the inputs to ground. This simple procedure permits you to check out the dc logic 0 and logic 1 operating conditions.

The R and S inputs are overriding. When they are used the unit operates as a simple R-S flip-flop. A basic flip-flop logic diagram is given in **fig. 2.** Generally when the set input is logic 1, the Q output is logic 1 and the $\overline{Q}$ is logic 0. In the case of the SN7472, there is an inversion in the S and R line as indicated by small circles in the logic diagram, fig. **2B.** This means that with input S set to logic 1, the Q output is logic 0 and the $\overline{Q}$ output is logic 1.

step 1:

Connect a voltmeter to the Q output. Leave the S terminal at logic 1 (no connection) and set the R terminal to logic 0 by connecting a short between the R terminal and ground. The Q output is zero and the $\overline{Q}$ output is positive logic 1 voltage (positive logic device). Reverse the logics of inputs R and S. Note the output logics flip over. What are the output logics with both the R and S terminals at logic 0? Do the same with both the R and S inputs at logic 1. The output logic for the latter condition can be 1,0 or 0,1 depending upon the previous logic of the R and S inputs. The above logic conditions hold regardless of the logic at the J, K and C inputs; R and S are overriding.

step 2:

Momentarily apply logic 0 to the R input and logic 1 to the S input to

establish the *set* state of the Q and $\bar{Q}$ outputs with $\bar{Q}$ at logic 1. Keep the R and S inputs on logic 1 (no connections). Set the K input on logic 0 and the J input on logic 1. Momentarily apply logic 0 to the clock input. Note that the output logic changes and holds whether the clock input is kept on zero or one. Momentarily apply logic 0 to the R input. This returns the output logic to its set state.

Now connect logic 0 to J and logic 1 to K. Again, momentarily apply logic 0 to the clock input. Note that the output logic again flips and holds. Momentarily apply logic 0 to the S input. Note that the output logic goes back to its reset state.

signal operation

Additional understanding of the operation of the universal flip-flop can be gained by using a signal from an audio sine- or square-wave generator and the 100-kHz crystal calibrator detailed in last month's column.

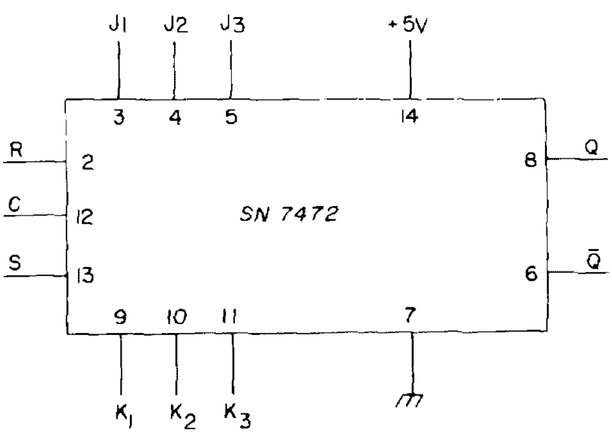


fig. 1. Connection diagram for SN7471.

step 3:

Set all inputs to logic 1 (no connection). Apply the 100-kHz output of the crystal calibrator to the clock input. Observe the signal at the Q and $\bar{Q}$ outputs. Compare the input and output frequencies. The unit is operating as a two-to-one counter. Disconnect the 100-kHz clock.

step 4:

Apply the output of the audio generator across the R-S inputs. To activate the unit it may be necessary to increase the audio gain above that required at the clock input. Output and input frequencies are now the same, but the waveform has been improved and the output is steep-sided compared to the input signal.

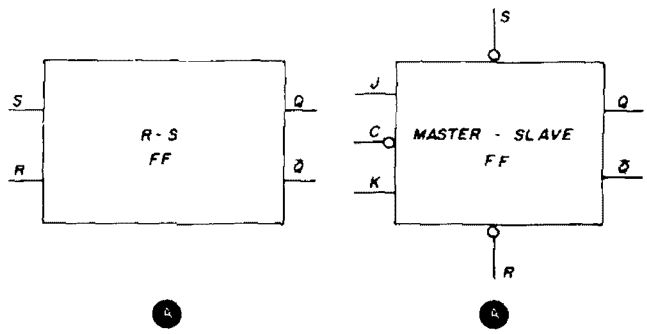


fig. 2. Basic and universal flip-flop.

step 5:

Disconnect the audio generator. Connect the output of the 100-kHz crystal calibrator to the clock input. Observe the output waveforms at Q and $\bar{Q}$.

Connect the output of the audio oscillator to the J input of the flip-flop. Set the audio oscillator frequency to 1000 Hz. Display one cycle of the 1000-Hz repetition rate on the oscilloscope screen. During approximately one-half of its period, the clock signal appears in the output while the second alternation contains no clock *component.* Momentarily connect 0 logic to the R input. Output is removed and the flip-flop is returned to the set state.

inhibiting

In an inhibiting circuit a certain logic at one of the input circuits prevents or inhibits the transfer of logic information between another input and the output. A basic inhibiting circuit is shown in **fig. 3.**

The transistor is normally non-conducting because the *emitter junction* is

not forward biased. This represents logic zero condition at the base. When a positive pulse of sufficient magnitude (logic 1 level) is applied to the base, the transistor conducts, and its inversion produces a logic 0 output. This is true provided no logic 1 voltage is applied to the emitter.

If a logic 1 pulse arrives at the emitter at the same time a logic 1 pulse is applied to the base, the transistor remains non-conducting. Therefore, the output of the transistor remains normal at logic 1 potential.

Likewise, the two other possible conditions of X and Y both being at logic 0 or both being at logic 1 result in a logic 1 output. The truth chart for the circuit is given with **fig. 3.**

In considering the universal flip-flop you learned that the R and S inputs were overriding. One might state that these inputs inhibit the clock's, J and K inputs just as input Y in **fig. 3** is able to inhibit input X. The inhibiting process is used to advantage in counters because they permit a binary counter chain (all even count) to also function in an odd-count manner. In our next experiment a divide-by-eight counter consisting of three binary 2-to-1 counters in series is made to divide by 5. This is done by feeding back inhibiting information from the output.

even and odd counting

The final construction experiment last month consisted of a crystal-controlled multivibrator and a 7472 two-to-one counter. The versatility of this calibrator can be extended with the addition of still another integrated circuit. In this case it is the inexpensive 7473 dual J-K master-slave flip-flop. Two binary counters are included in the same case with separate inputs, outputs and clear terminals, **fig. 4.** They can be operated separately or joined together to obtain an overall count of four-to-one. Combined with the previous counter, a total division of eight (2 x 2 x 2) can be obtained.

X	Y	Z
1	0	0
0	0	1
1	1	1
0	1	1

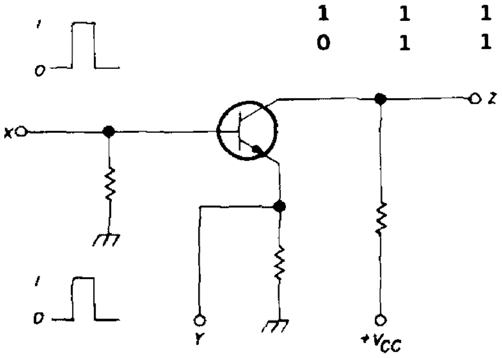


fig. 3. A basic inhibitor circuit and its truth table.

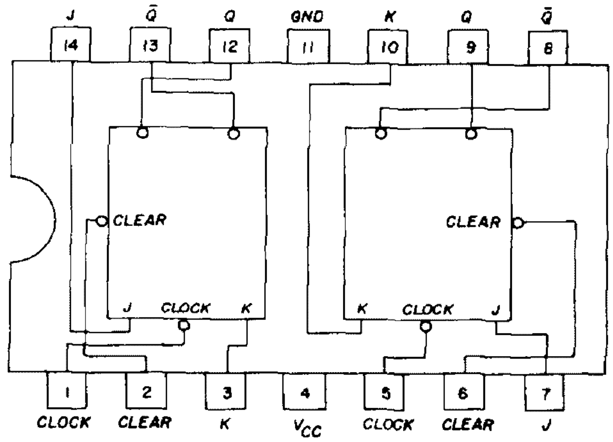


fig. 4. The basic 7473 dual flip-flop containing two binary counters.

step 1:

Connect the circuit as shown in **fig. 5.** The output of the previous counter is joined to the new dual flip-flop by connecting its $\bar{Q}$ output to the clock input, pin 1. The two counters in the 7473 are joined by tying together the $\bar{Q}$ output at pin 13 to the second clock input at pin 5. A divide-by-eight output is made available at either pin 8 or pin 9.

step 2:

Turn on the calibrator circuit, connect the oscilloscope to the output of the crystal oscillator and display eight cycles on the scope screen. Now connect the *oscilloscope to* the output of the last counter. Note that only a single waveform is displayed, indicating a division by eight. Two cycles will be seen at the junction between the two dual counters and four cycles at the output of the first

counter. Waveforms will be those of **fig. 5.**

step 3:

Now connect the circuit of **fig. 6**. A 2000-pF capacitor is used to connect the Q output of the last counter to the reset or clear inputs of the first and second counters. Capacitor C1 provides a feedback path for a reset pulse. Pulse polarity is such that it cancels out or inhibits the activities of the first two counters at the proper time to start a new five-pulse sequence.

Note in **fig. 5** (divide-by-eight counter) that the edges of all five waveforms are coincident at the leading edge of the fifth clock pulse. Furthermore, the trailing edge of the Q output of the last counter is swinging from logic 1 to logic 0, a polarity which can be used to clear and reset the first two counters. This very activity is shown as coinciding with the leading edge of the fifth clock pulse in **fig. 6**. The leading edge of the sixth clock pulse will now switch all three counters just as the leading edge of clock pulse one at the very beginning. The counters see everything as starting anew and they go off to try and count eight again only to be met by another reset pulse coincident with the leading edge of clock pulse ten.

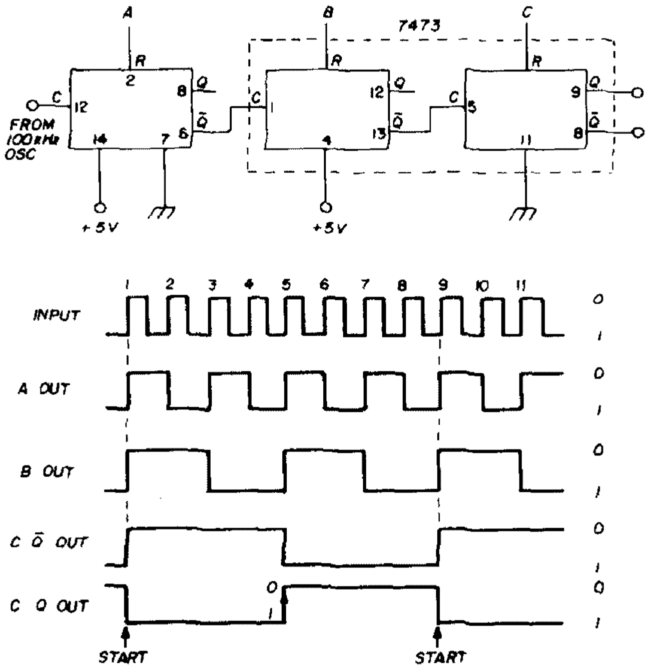


fig. 5. An 8-to-1 divider and corresponding waveforms.

The feedback spikes shown in the counter 1 and counter 2 outputs coincide with the leading edge of the fifth clock pulse. Observe these spikes on the oscilloscope. Note how the entire cycle of events repeats itself. Actually, the counter chain is never permitted to go through its eight-count cycle but is interrupted in a manner that continues a five-count sequence.

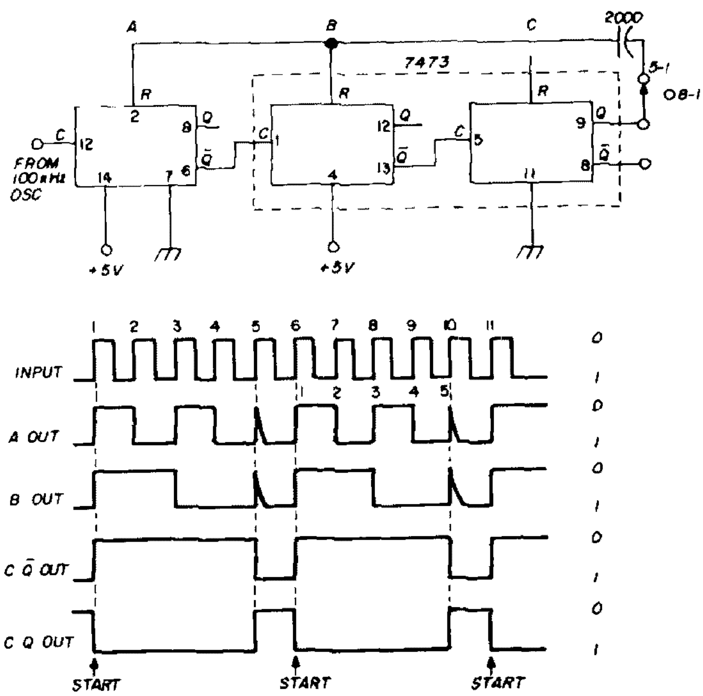


fig. 6. A 5-to-1 divider and corresponding waveforms.

step 4:

Turn on the counter and observe the waveforms at the outputs of each counter. Look for the inhibiting spike in the outputs of the first and second counters, exactly as shown in the waveforms of **fig. 6**.

step 5:

Attach the oscilloscope to the output of the last counter and adjust it until eight cycles are displayed. Note also that the output is now asymmetrical (pulse of shorter duration then the intervening spacing). Recall that the outputs for the even-count activity were symmetrical square waves. Momentarily disconnect

the capacitor from the Q output of the last counter. How many cycles are displayed now? The number has dropped to five, indicating a change back to an even eight-to-one count.

recommended by H. Beverage in his original paper. The correction to the 200-ohm terminator can be multiplied by quite a large factor due to the somewhat non-linear instantaneous current distri-

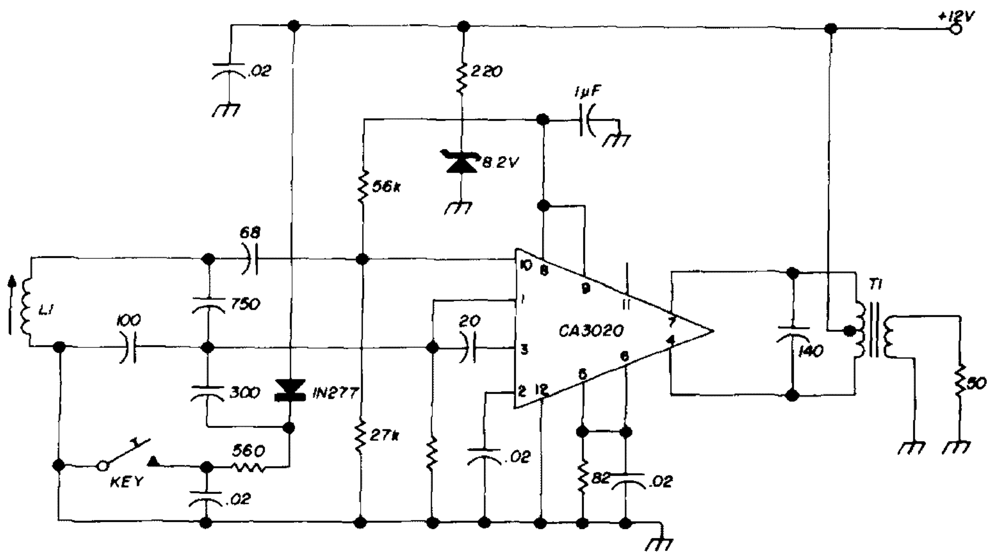


fig. 7. Seiler vfo with CA3020 developed by W9ZTK. T1 is a toroidal transformer.

step 6:

Operate the counter in the divide-by-five mode. Connect the output to the input of your receiver. As you tune over the band you will find a calibration point every 20 kHz. This pulsed output is again rich in harmonics with marks to be found over the 6-meter band, and even on 2 meters with proper low-loss coupling.

Beverage termination resistance

Here is an instructive note from Robert N. Morris, W7ALU. It has to do with the normal 200-ohm termination value for a Beverage antenna,

" . . . this is a close to theoretical value as per the radiation resistance of a non-resonant terminated antenna about three and one-half wavelengths long and being remote from earth. The Beverage however, not being remote from earth, has a very high earth loss for both transmitting and receiving. A much higher order of termination would be indicated as being necessary due to the very low height of the Beverage, ten to twenty feet being bution in any terminated wire. The value of radiation resistance may differ slightly from the characteristic impedance owing to the termination along the wire due to radiation in transmitting or losses in the wire in receiving. In these cases, the values for the feedline and the termination resistance would be different. Therefore, both input and output values should be raised or lowered. These values should be set somewhat toward those values necessary for the center wavelength to be used. Variations exist as to a particular height and location and according to the principal height of the electrical ground.

"Due to the high ground losses, the Beverage can never produce the high signal results as the same antenna if it were remote from earth, but all the same, the practically complete attenuation, (in the earth) under the Beverage of most *minor* lobes gives an apparent boost to received signals arriving off the terminated end."

keyed vfo

Cal Sondgeroth, W9ZTK, modified the

CA3020 QRP unit described in the August, 1971 issue of *ham radio.* His stable and clean keying circuit is shown in **fig. 7.** His description is as follows:

"... as you can see, I used a slug-tuned coil for L1 with the slug attached to a shaft for tuning from the front panel. Grounding the slug was important to eliminate hand-capacitance effects. The shaft is fitted with a little bearing near the panel, and this tuning arrangement works out to be very stable even though it's not calibrated.

"The main drawback of the oscillator at first was that it drifted quite badly every time it was turned on. Putting a soldering iron on the IC seemed to indicate that it was pulling the frequency as it warmed up. Attempts to let it run all the time and open the emitter lead of the oscillator transistor to shut off the oscillator really didn't seem to help much, and this doesn't provide good keying for CW either. So, I came up with the diode-keying circuit shown. This allows everything to be on all the time to level out the temperature of the IC. With the key open, oscillation is prevented by the fact that there is no capacitance (or 100 pF as in my circuit) from emitter to ground on the oscillator. Oscillations can be stopped in this way, or the oscillator can be shifted off the oscillating frequency by moving about 100 kHz during standby. The frequency shift can be varied by the amount of capacitance switched in and out of the circuit. For CW keying, of course, it would be necessary to completely stop the oscillator. Frequency-shift keying for RTTY can be obtained by the same sort of circuit with smaller values of shift capacitance.

"My output transformer is a toroid coil with bifilar primary. The secondary is matched to about 50 ohms, and I did not get as much output as I expected. However, this is a secondary consideration for a vfo, anyway. With the keying circuit shown and the zener regulator, the keying is now really stable and after all, stability was what I was after."

ham radio

the ham notebook

simple intercom

Here's a simple intercom that can be put together in a couple of nights and is guaranteed to pacify the wife when you're in the basement and dinner is waiting on the table. It uses a half-watt usually have a high-impedance output, this IC was made with a high-impedance input. The output, however, is low impedance and will work directly into an 8-ohm speaker without the usual output transformer.

The circuit, shown in **fig. 1**, is very

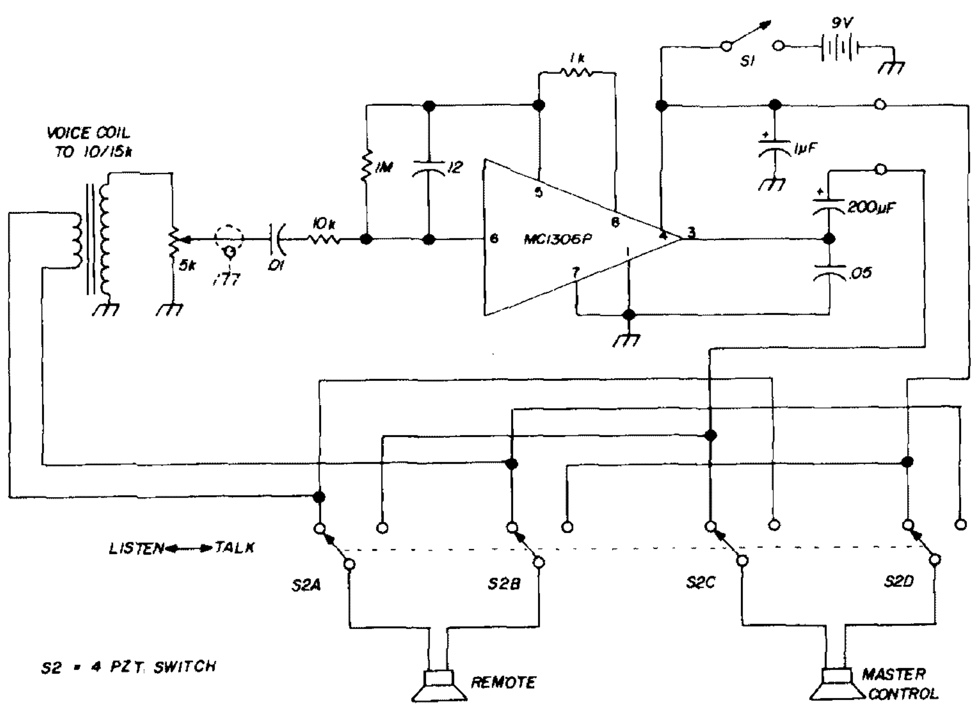


fig. 1. Schematic of the simple intercom.

audio IC, Motorola MC 1306P ($1.10). This IC was designed for use in portable a-m/fm radios and tape recorders and takes a regular 9-V transistor radio battery for power. Since these circuits simple. I made a small etched-circuit board for the IC and associated components. For the master control, I used an old ac-dc radio, with all parts removed execpt the speaker and its output trans-

former. The output transformer is necessary to step up the input (talking) speaker to a high impedance. The etched-circuit board and battery were mounted inside and the *talk* switch (Burstein Applebee 18A1309) was mounted in the dial hole with the volume control in the regular volume control hole. Shielded cable from the volume control to the input may be necessary. The remote can be any 8-ohm speaker in whatever enclosure is available.

Once installed, the remote is on all the time and any activity or talk can be heard by the control. By pushing the switch to the *talk* position, the master control can talk to the remote. In my setup the remote picked up so much noise from machinery that it was distracting. To remedy this I ran a 4-wire cable instead of the usual 2-wire. I used the two extra wires for a buzzer at the control end and a push-button switch at the remote end. I

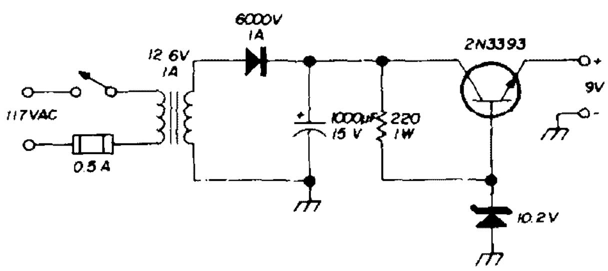


fig. 2. Power supply to operate the intercom from a 117 Vac source.

now leave the unit off (unless I want to listen in or call the remote) and the remote can buzz when someone wants to talk. Since there are no tubes to heat up, the unit will work as soon as it is turned on. This also conserves battery power. An ac supply can be used, and one is shown in fig. 2. This will probably have to be mounted outboard to avoid hum pickup from proximity to the circuit board.

When this unit was first turned on it picked up the local broadcasting station, but a 0.002 bypass capacitor from one side of the remote line to ground cured the trouble. In extreme cases of this type, try different values of capacitors on either or both sides of the line to ground. Perhaps chokes in both lines would help.

The IC will take up to 12 Vdc, but more than 9 V will make it run hot and

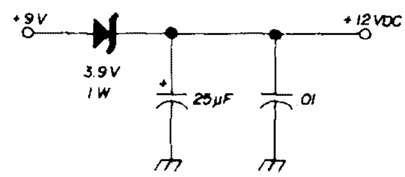


fig. 3. Voltage dropping and regulating circuit to operate the intercom on 12 Vdc.

high voltage isn't really necessary. If you have 12 Vdc available and want to get 9 V, try the circuit in fig. 3 using a 3 V, 1 W zener. A dropping resistor will not work as the current varies with speech input and a stable 9 V would not result. Also note that neither side of the remote line is grounded because the positive voltage is on the line for the output speaker.

Nat Stinnette, W4AYV

miniature power supply transformers

When powering really compact equipment such as electronic keyers and digital devices, obtaining a suitable tiny power transformer may prove difficult and expensive. This fact came to light while I was designing a sixteen IC keyer which was to be no larger than a conventional bug, yet completely self-contained including squeeze paddles, monitor and power supply.

The requirement for a miniature *power* transformer was met by using an audio output transformer intended for service in pocket transistor radios. A bridge rectifier of small glass diodes across the voice coil winding and a simple resistor-capacitor filter provided a full 3 V at 200 mA. The compact keyer, made possible by the tiny transformer, has been in service for four years without failure.

Gene Brizendine, W4ATE

meter safety

Amateurs often overlook the fact that high-voltage meter multipliers are dependent upon the meter coil for their return to ground. Should this coil open, the *entire high* voltage will appear at one terminal of the meter! To preclude this risk, a very high value resistor, Rx, is connected across the meter as shown in **fig. 1.** This resistor must be large enough to introduce only a small error in the meter reading, which may be compensated by a

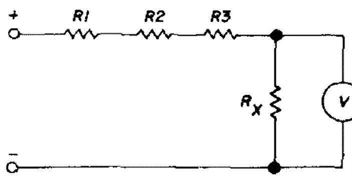


fig. 4. Use of a shunt resistor for safety.

corresponding decrease in the value of R2 to correct the meter.

As a further note of safety, the multipliers, R1 through R3, generally should be 2 W or larger, because a 500 V maximum drop across each resistor is an acceptable value. Otherwise, the resistor drift due to heating may be excessive and the voltage breakdown of the multipliers will be exceeded.

M. H. Gonsior, W6VFR

electronic fence interference

I live in a built-up area which permits keeping horses. I have had very strong noise interference on one bearing on the higher bands, and on all 3.5- and 7-MHz signals when I use a nondirectional antenna. I found the source of the noise very quickly with an inexpensive vhf aircraft-band portable a-m radio.

The interference consists of about one buzz per second, somewhat more regular than thermostatic devices. It is caused by the less expensive *Sears weed and stock-control charger with pulser timer,* which is described more extensively in the *Suburban and Farm Catalog* than in the general catalog.

Bill Nelson, the radio interference expert with The Southern California Edison Company and author of a very good article on the subject several years ago in *QST,* says that some electric fences produce only a series of clicks, while others have the buzzes.

During a 21-MHz contact with VK3AKB, Bill turned on the charger that was located near his ssb equipment and heard no interference at all.

The Sears general catalog did not mention any radio interference filter in the lower-priced charger, but did for the more expensive one. The farm catalog, in addition to mentioning the filter in the $40 unit, says that the $24 unit will not interfere with radio and tv. The noisy one here puts intermittent show on a tv screen near the noise source, and can be heard a hundred yards away on 140 MHz.

Sears has written extensively to assist in eliminating the radio interference and has provided factory comment. They say that both of their chargers are shielded and filtered to prevent radio, tv and telephone interference. When the interference is present, it is usually caused by one or both of the following conditions:

First: A current leak to ground somewhere along the fence line. This could be caused by very dirty or broken insulators, the wire touching against the side of a post, tree or building or heavy vegetation growing up against the wire.

Second: A loose connection somewhere along the fence line. This could be due to poor splices (usually just twisted connections in a rusting wire), a gate opening or very badly rusted spots in the wire.

Either of the two conditions above would cause a gap across which sparks could jump, exciting the fence wire as an antenna, thus resulting in very broad interference to radio, television or telephone. A careful check of the fence line should disclose the trouble — and may require some soldering for permanent connections.

Sears also says that you can determine definitely whether the controller (charger) itself is at fault by disconnecting it

from the fence and allowing it to operate. Of course, the removal of the antenna will reduce the interference markedly in any event. However, if a nearby receiver shows that the interference disappears entirely, the fence rather than the charger is at fault.

If the noise continues, the charger unit – possibly due to an open filter condenser – is at fault and may be returned to a service station for repair. First, however, you may try a new pulser in the charger, as this possibly could be the source of the difficulty.

E.H. Conklin, K6KA

low-band vertical antenna

The 4BTV four-band vertical antenna requires frequent readjustment when you are changing frequency and using the 75-meter resonator or loading coil. The same problem is present when using base loading. If a remote motor drive is used, additional cabling is required. There is a solution available that allows the operator to tune the 4BTV over the entire 80- and 75-meter band from his station location and still maintain a reasonable bandwidth. The technique uses a coaxial quarter wave transformer tuned to 3.750 MHz (43.3 feet long) and a broadcast type ganged variable capacitor connected at the remote end as shown in **fig. 1**.

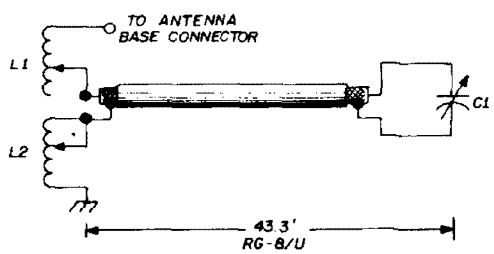
In my case, the tap on L2 for proper impedance matching occurs at the top of the coil. L1 requires exactly half of the 30 available turns when C1 is half meshed. The vswr is 1:1 at the mid-band resonance frequency of 3.750 MHz. The inductive loading for quarter-wave resonance and a near 50-ohm resistive impedance is found by varying the tap on L2 while L1 is varied for quarter-wave resonance (L1 and L2 are mounted with zero mutual coupling).

After L1 and L2 are adjusted for a minimum vswr at 3.750 MHz with C1 at half mesh, then C1 need only be adjusted at any other frequency in the 80-meter band for a vswr better than 1.3:1.

The arrangement and components used have been exposed to a 300-watt power output level. At 500-watts output, the base antenna connection of the 4BTV arced across to ground. Therefore, 30-watts output is used with the 4BTV on 80 meters with this arrangement.

The stub may lie on the ground since it is connected at the low-impedance point (shield of coax to top of L2 and center conductor to bottom of L1). Negligible loss and capacitance-to-ground effects are experienced and the quarter-wave transformer inversion properties present a variable series inducttance and capacitance between L1 and L2. The coils, once set at 3.750 MHz, require no further adjustment over the band.

A 4BTV can be replaced by a full quarter-wavelength 40-meter vertical or an all-band vertical with equally good



L1	30 turns, no. 12 wire, 8 tpi, 2.5-inch diameter
L2	6 turns, no. 12 wire, 2.5-inch diameter. Turns spaced approx. 5/8- inch apart
C1	five gang broadcast type, 365 pF per section

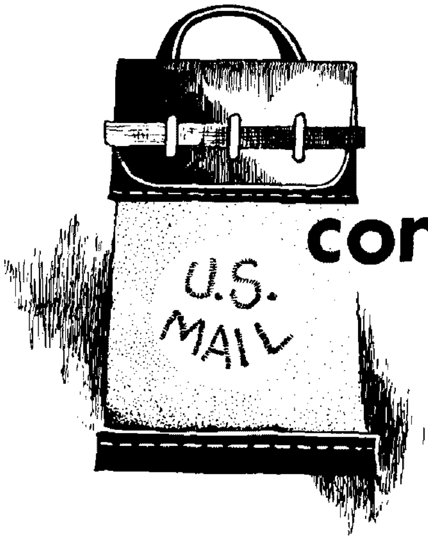
fig. 5. Remote antenna tuning arrangement.

performance on the 80- and 75-meter bands.

One word of caution: When using base loading of a quarter wavelength 40-meter vertical antenna on 80 meters, a healthy rf voltage will appear at the base of the antenna, a hazard to unsuspecting people or animals.

The values of L1 and L2 may vary at different installations and must be determined experimentally. Those given in fig. 1 should be about right for those installations using one ground rod and mounting pipe with the 4BTV.

Fred M. Griffee, W4IYB



comments

speech clipping

Dear HR:

I write this letter to question the general understanding and use of rf clipping circuits as given in many schematic diagrams; namely, I question the conventional explanation given for clipping at an intermediate frequency such as 9 MHz.

Squires and Bedrosian ("The Computation of Single-Sideband Peak Power," *Proceedings of the IRE,* page 123, January, 1960) have used a Fourier analysis, coupled with the concept of "frequency incommensurability" to show that audio clipping results in an *increase* in ssb peak amplitude rather than the expected reduction. This, of course, arises because a single-sideband signal does not preserve relative phase information; hence, there always exists a point in the ssb envelope where all frequency components add for an instant to produce a peak (the more frequency components present, the higher the peak). Thus, because audio clipping produces additional frequency components, the rf peaks become more intense, contrary to expectations.

However, the mathematics of the subject article also appear to be applicable to a situation where one clipped rf wave modulates another rf sine wave (heterodyning), and one of the sidebands (say, the difference frequency) is rejected. In this case, as in the audio case (which is not really a special one), phase information is lost, and one can use Squires' and Bedrosian's mathematics to show that the final signal should be degraded too. I therefore feel that this is a matter which ought to be clarified, because this particular article has become the *raison d'etre* for rf clipping.

It appears that if Squires' and Bedrosian's analysis is correct, then rf clipping can only be expected to produce optimum results when it is used at the *operating* frequency. If that is true, as my own brief inspection seems to indicate, then someone who has the time to pursue it further should discuss in detail this apparent negation of rf clipping. After all, how do you filter the out-of-passband products at the operating frequency?

Furthermore, if my interpretation of the mathematics is corrent, clipped dsb should be far more effective than unclipped dsb, whereas clipped ssb will supposedly be a disappointment.

Oscillograms appearing in the 1969 "Radio Amateur's Handbook" appear to support the notion that rf clipping at intermediate frequencies works, but one can again take note that "spikes" (many

of them) appear in the output of a clipped and heterodyned, ssb waveform, and that they do not follow the i-f envelope faithfully, all in accord with the subject article's predictions. Perhaps clipping at the i-f is effective, but if it is, someone must look more closely for the reasons and, perhaps, determine if Squires and Bedrosian produced an analysis that was too simplified. (It never really concerned itself with speech — merely distorted "sine" waves — and it totally ignored filtering.) As a matter of fact, I suspect that the analysis *is* over-simplified and that amateurs may have succeeded in regions not really covered by theory. Perhaps there are later papers which clarify the situation, and one of your readers who keeps up with IEEE's journals can help out. I hope so, because in my mind the case for conventional rf clipping does not seem to be on solid ground.

Richard R. Slater, W3EJD

Dear HR:

I was pleased to get the reference to the Proceedings of the IRE note by Squires and Bedrosian, since it nicely supports my simplified explanation of why speech clipping and the ssb mode are incompatible. (See my article in HR *for February, 1971) I was not aware of the material though I have the Proceedings issue in my collection!*

Let me start by assuring you that rf (or i-f) clipping is indeed well understood. I believe your difficulty arises from the fact that in the IRE article an infinite bandwidth is assumed. This is why the authors get even worse results for clipped speech in a ssb system than I, because I assumed practical limits for bandwidth (3 kHz) and the lowest audio frequency (400 Hz). An infinite, or to be more practical, a large bandwidth, is a purely relative term and must be viewed against the signal frequency. A 3-kHz bandwidth is indeed nearly infinite when you consider the distortion products or harmonics of a clipped 300-Hz tone. However, a 3-kHz or even a 30-kHz bandwidth is small when you consider a 100-kHz (or 9-MHz) clipped signal.

Since the distortion products or harmonics (multiples of the frequency of the clipped signal) are sufficiently filtered out by a single i-f transformer after the i-f clipper, the problem which is causing your concern does not arise. The subsequent mixers and amplifiers are dealing with an amplitude-limited signal without *the distortion components.*

Some time ago, there was a widespread belief that when you filtered a clipped signal so that all harmonics were removed, you got back the unclipped original. This of course is a fallacy; inspection of the Fourier series for a square wave shows that the output variation is 2 dB for inputs between 1 (the clipping level) and infinity!

Within the context of your letter, clipping at audio is a special case, since the lower harmonics generated by the clipper fall into the band of interest. Clipping at i-f or rf avoids this through the selectivity inherent in most designs so that there is no phase information to lose. (I agree of course that later mixers are in effect ssb mixers.) For example, when clipping a 300-Hz tone, the distortion products are at 900, 1500, 2100, 2700 Hz etc. When you clip a 100-kHz ssb signal, the distortion terms show up at 300 kHz, 500 kHz etc. Obviously, it is no major task to get rid of these, as the band of interest is still only 3 kHz wide.

Walter Schreuer, K1YZW
Ipswich, Massachusetts

mosfets maligned

Dear HR:

An article in the March, 1972 issue, Gerald Vogt's writeup on a two-meter preamplifier, inspires this missive, since I am of the opinion that he leaves some misrepresentations in his article.

I am pleased to see that Mr. Vogt

recognizes the ease with which the classic cascode amplifier circuit can be directly implemented with fet devices. The excellent results that may be obtained with this circuit arrangement without the need for neutralization are recognized advantages of the cascode amplifier. The 2N4416 or other jfets can produce a fine two-meter preamplifier.

However, I do feel that the author has been inclined to dismiss the dual-gate mosfet devices in much too casual a manner, and without a sufficient understanding of their true virtues. The argument that the mos junctions are subject to destruction by electrostatic discharges is archaic in this day of the diode-protected device. Certainly the use of dual-gate mosfets in the Heath SB-303 shows the practicality of the protected device. I am presently involved with a receiver design which is in production, using RCA dual-gate 40822 fets, and we have used some 4000 devices on the assembly floor without a failure due to electrostatic problems. Incidentally, I am using this type fet for rf, mixer, i-f, audio and oscillator stages and find it admirably suited for all applications.

The inherent cascode internal connection of the dual-gate mosfet makes it ideally suited for cascode preamplifiers, and it is regrettable that its superiority was not recognized.

The dual-gate device offers better intermod and crossmod performance than any other device in the solid state museum, and permits the introduction of an rf gain control or agc function that actually *increases* the signal handling capabilities as the gain is reduced. This is even better performance than that offered by the pentode variable-mu vacuum tube!

I would suggest that Mr. Vogt read over some of the informative application notes produced by Motorola and RCA, for example, on the use of dual-gate mosfets in rf applications. Data on the RCA 40673, RCA 40822, or Motorola MFE121, for instance, will reveal that these devices will give noise figures in the 2- to 5-dB range at two meters with stage gains near 20 dB as cascode amplifiers at 150 MHz.

RCA application notes give particular insight into uses of these fets. RCA publication ST-3233 on small-signal rf amplification of mos devices; AN-3435 on cross modulation effects; ST-3486 on receiver applications of dual gate mosfets and AN-4431 on rf applications of the dual-gate mosfet up to 500 MHz should be particularly informative.

Finally, I am surprised that Vogt's circuit does not include any local feedback, such as a source resistor with suitable rf-bypass capacity. Since the parameters of fets vary considerably from device to device, it is very important to incorporate dc feedback to make the circuit less critical of device parameter controls. Since Idss varies from device to device, and Gm does also, it is better to smooth out these variables with feedback than to accept large performance variations in the circuit as devices are changed.

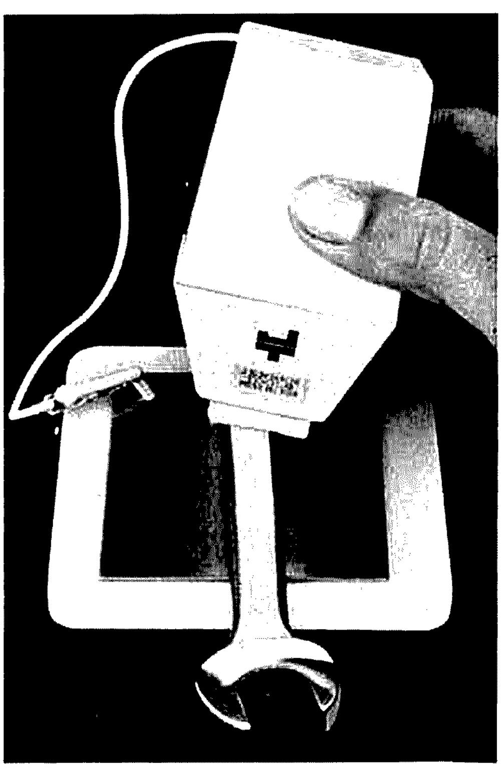
I fear that the circuit given by Mr. Vogt will function best only with low-level signals, since nothing is done to optimize the large-signal capability. Further, an rf gain control can be readily introduced to a dual-gate cascode circuit by controlling the voltage applied to gate number 2. If this gate is biased initially at say 30 to 40 percent of the (drain) supply voltage, rf gain can be easily reduced by merely lowering this gate number 2 voltage, which at the same time actually *increases* the signal-handling ability while reducing the stage gain. Certaintly this is a virtue not to be so lightly dismissed.

The above comments are not intended to be hyper-critical of Mr. Vogt's article, as certainly the use of fets as a cascode rf amplifier is far superior to usual bipolar circuits. However, I do feel that the mosfet, especially in the diode-protected dual-gate form, should not be overlooked as the best of the solid-state devices for small-signal circuit designs.

Maurice P. Johnson, W3TRR
Randallstown, Maryland

new products

metal-etching tool



You can mark tools, metal components, chassis as well as camping gear and household items with a new metal-etching instrument from Jensen Tools. The Model 17 Marking system can electronically etch flat or round metal surfaces and can make two-color metal labels. The tool is safe, silent, vibration free and powered by ordinary 117 Vac.

Simple to use, all that is required is to type, write or draw the mark to be etched on the special stencil included in the kit. A wetting solution is applied to the head of the marker, stencil attached, and the tool is ready for impression making. Chrome plated and solid brass labels with strong adhesive backing are available for etching with a black, copper, or brass-colored mark for a striking two-tone effect.

Model 17 is packaged in an economical kit which includes the marking tool, electrical cord, ground plate, electrolyte solution in plastic bottle, an adapter clip for deep etching and full instructions. The kit includes supplies sufficient for marking up to 2000 items. It is priced at $24.95. Refill supplies are readily available.

Further information is available by using *check-off* on page 110 or by writing to the manufacturer and requesting the descriptive folder on the Model 17 Marking system from Jensen Tools and Alloys, 4117 North 44th Street, Phoenix, Arizona 85018.

great circle map

A new edition of the Radio Amateur's Great Circle Chart of the World has just been published by the Radio Amateur Call-book. Measuring 29" by 25", the six-color map is centered on the geographical center of the United States and includes a chart giving great circle bearings to all parts of the world from Boston, Miami, Seattle, Los Angeles, San Francisco and Washington, DC. The chart, an azimuthal equidistant projection, shows the great circle course from its center point to any other point on the earth as a straight line. Additionally, distances along the straight line can be

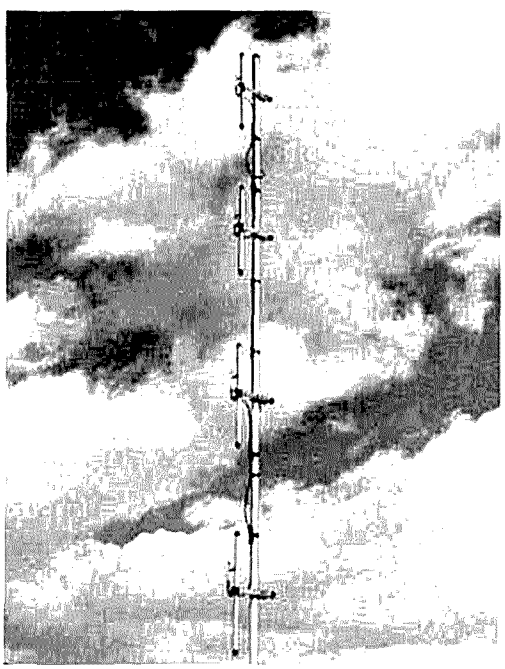
More Details? CHECK—OFF Page 110

measured accurately against one standard scale.

The chart, besides dressing up a radio shack, is helpful for orienting antennas and comparing long and short paths to distant spots. The map amazes visitors to the station as it shows some interesting facts such as the shortest path to Singapore, Vietnam or Burma from Greenville, New Hampshire right over the North Pole.

The map is available for $1.25 from Comtec Books, Greenville, New Hampshire 03048.

four pole antenna



Cush Craft announced the addition of two new models to their Four Pole antenna design. The Four Pole is a series of four stacked dipoles for amateur fm and commercial use.

Four Poles are supplied with the dipoles, mounting booms, harness and all hardware. The center support mast is not supplied, allowing the user to custom select a mast for his installation or to tower mount the antenna.

Gain figures for the antennas show 6-dB omnidirectional and 9-dB semidirectional pattern. The three models now available are the AFM-4D, 144 - 150 MHz, $42.50; the AFM-24D, 220 - 225 MHz, $40.50 and the AFM-44D, 435 - 450 MHz, $38.50.

More information is available from Cush Craft Corporation, 621 Hayward Street, Manchester, New Hampshire 03103 or from *check-off* on page 110.

circuit zaps

Circuit Zaps, copper component patterns and accessories used to produce instant printed-circuit boards, have been developed for the hobbyist. International Rectifier Corporation recently introduced its complete line of Circuit Zaps for custom and prototype production of printed-circuit boards.

Circuit Zaps, which will retail for as low as 9 cents per pattern, enable the hobbyist or design engineer to eliminate the artwork, photography, photoprinting, touchup, etching, stripping and other time consuming and costly steps previously required in prototype and test circuit development and in home electronics projects.

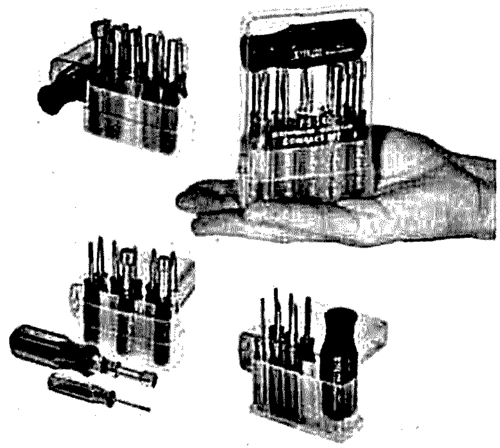
Circuit Zaps are available in four design groups, and each card contains three to twelve of one pattern. Each pattern is precision-etched on a 5-mil glass-epoxy base material and backed with a special pressure-sensitive adhesive.

In actual use, the Circuit Zaps are placed on a pre-punched or unpunched circuit board in any desired layout. The Circuit Zaps can be removed and repositioned without damage to the adhesive.

The board is then ready for testing, drilling (for unpunched boards) and positioning of conductor paths. The final step is the mounting of components with standard hand-soldering techniques.

Specific Circuit Zap patterns available include: TO-type patterns with 3, 4, 6, 8, 10 and 12 leads; dual in-line with 14 and 16 leads; resistor/diode types, conductor path and both single-and dual-component pads. Also available from the Semiconductor Division of International Rectifier are: unpunched XXXP laminates, prepunched with 0.1 x 1.0-inch centers and printed-circuit board terminals. Available at your local dealer. For more information use *check-off* on page 110.

compact driver sets



Xcelite has just added three new all-purpose screwdriver and nutdriver sets to its family of "compact convertibles." Each set consists of an assortment of color-coded midget tools and a unique "piggyback" torque amplifier handle which enlarges gripping surface, extends reach and increases driving power. The new units bring to nine the number of "compact convertible" sets now available, with various assortments of drivers for slotted, Phillips, Allen, Scrulox, hex and clutch head screws, plus hex nuts.

Featured is a new transparent container with a positive snap-lock. Optically clear for easy set identification, the tough, injection-molded cover stays closed even when tossed into a tool box. The case is also designed to hold tools upright on a bench for easy selection. All nine members of the "compact convertible" family are now housed in these transparent, "show case" cases.

Contents of the three new sets are: PS-6 Screwdriver Set – miniature drivers for No. 00, 0 and 1 Phillips; 3/32", 1/8" and 5/32" slotted screws.

PS-140 Screwdriver and Nutdriver Set – popular assortment includes drivers for No. 0, 1, and 2 Phillips screws; 3/32", 1/8", 3/16" and 1/4" slotted screws; and 1/4", 5/16" and 3/8" hex nuts.

PS-130 Screwdriver and Nutdriver Set – similar to PS-140 with larger assortment of nutdrivers, 3/16", 1/4", 5/16", 11/32" and 3/8" hex sizes; plus drivers for No. 1 and 2 Phillips screws, and 1/8",

3/16" and 1/4" slotted screws. Both this set and PS-140 are expansions of the very popular PS-7.

Complete details are given in the new Catalog 171 available without charge from Xcelite Incorporated, Orchard Park, New York 14127 or by using *check-off* on page 110.

current-controlled resistors

Radiation Devices has introduced two new current-controlled resistor (CCR) pairs. CCR pairs and quads may be used in tuning, switching, attenuating or isolating functions in electronic equipment or systems. They consist of a light-emitting diode optically coupled to a pair or quad of cadmium selenide photo-resistors, selected and adjusted so that their resistance values track over a thousand-to-one range. The control current through the LED required to cover this range varies from a few tens of microamperes at one megohm to 20 milliamperes maximum at one kilohm. Fourteen pin DIP packaging, combined with carefully considered pin assignment, provides convenience in utilization, compatibility with modern circuit-board technique and maximum signal isolation.

The model 500-104 is a resistor pair selling for $18, and the model 500-105 is a resistor quad selling for $24. Complete specifications are available by using *check-off* on page 110 or by writing to Radiation Devices Company, Box 8450, Baltimore, Maryland 21234.

logic tester

Requiring no visual observation, the model 95 logic tester gives a high audio tone to indicate logic 1 state and a low audio tone to indicate a logic 0 state when checking normal 5 V digital logic circuits. The unit sells for $19.95. More information is available from Production Devices, 7857 Raytheon Road, San Diego, California 92111 or by using *check-off* on page 110.

More Details? CHECK—OFF Page 110

high-current logic driver

High-current loads can now be driven by logic circuitry, using the Motorola MCH2890, dual power driver rather than using more complex discrete circuitry. This new device translates logic voltage levels to high-power outputs. Either DTL or TTL logic levels may be used to control the device, and loads can be either resistive or inductive.

Many applications such as RTTY magnet drivers and tape punches, hammer-drivers, relay drivers, stepping motors and lamp drivers require high current pulses that are digitally controlled. The new device provides this interface in a single package replacing an IC and two Darlington transistor packages.

The MCH 2890 combines a dual 2-input MTTL AND gate similar to the MC3101 and a pair of Darlington power transistors in a hydrid design to provide up to 6 amps at 10% duty cycle and 25 ms pulse width. Continuous output current is 1 ampere maximum. The output Darlington transistors have 120 V minimum breakdown voltage ratings which is desirable for driving inductive loads at high current.

A factor which has hampered IC drivers in the past was package power dissipation. A new 10-pin aluminum package similar to the popular TO-3 was designed for the MCH2890. Besides the power handling capabilities of the TO-3 package, it was also chosen because of its longtime popularity as the standard industrial power package.

For further information contact the Technical Information Center, Motorola Inc., Semiconductor Products Division, P.O. Box 20924, Phoenix, Arizona 85036 or by using *check-off* on page 110.

itu lists available

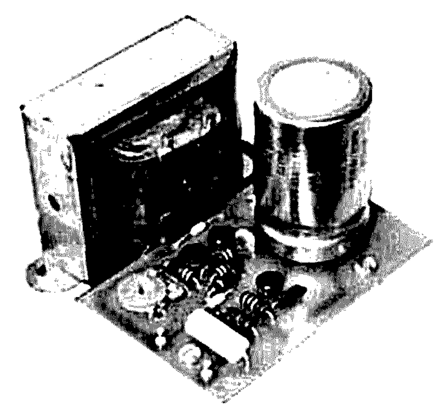
Formerly available only from Geneva, Switzerland; International Telecommunications Union publications are now avail-

able in this country from Gilfer Associates. The ITU, world-wide treaty registration center for all radio stations and frequencies, maintains computerized lists of the hundreds of thousands of radio stations. The lists are updated quarterly.

Typical lists and prices are: "List of Ship Stations" ($5.25); "List of Coast Stations" ($6.60); "List of Fixed Stations ($22.00); "List of Broadcasting Stations Operating Below 5950 kHz" ($11.50) and "Alphabetical List of Call Signs" ($7.70). Gilfer will also accept subscriptions to the ITU monthly magazine, "Telecommunications Journal."

A list of ITU publications and prices is available from *check-off* on page 110 or directly from Gilfer Associates, Inc., Box 239, Park Ridge, New Jersey 07656.

ic power supplies



Viking Electronics has introduced a line of low-cost power supplies for logic and linear-system applications. Priced from $17 to $24 in single quantities, the OEM 70 series provides typical outputs of 3.5 to 6 V at 3 A and 8 to 15 V at 1.2 A with regulation of 0.5 to 0.1 percent and ripple of 1 to 2 mV, dependent on models.

Features include electronic current limiting, floating output, stable differential amplifier circuitry, silicon transistors and computer grade capacitors. For more information contact Viking Electronics, Inc., 721 Saint Croix, Hudson, Wisconsin 54016 or use *check-off* on page 110.

More Details? CHECK—OFF Page 110

(continued from page 4)

amateur gear, plus all the possible mixes of sums and differences, you can appreciate the magnitude of the problem.

Recently, a well known fm'er prevailed upon an airline to test his Motorola HT in one of their aircraft so he could operate during a flight he planned to take. After months of correspondence and personal appearances at airline headquarters, and meetings with communications department people, many of whom are amateurs, the airline agreed to run the necessary tests.

The fm'er tweeked and peaked his trusty HT and checked it thoroughly with a spectrum analyzer; it was clean. On the appointed day the aircraft was removed from line operation and the test began. The test required three hours and four men to complete. The HT caused absolutely no interference, and the fm'er received a letter authorizing the operation of *that* HT on *that* particular trip in only *that* type of aircraft. It is easy to understand why the airlines, who are trying to cut costs, would prefer not to get involved in testing each amateur's fm rig.

Unfortunately for the fm'er who went to all this trouble, the aircraft on which the tests were conducted is to be phased out of operation soon, and he's right back with the rest of us — speechless while aboard an airliner.

Many fm'ers continue to ask the Captain's permission to operate, and he may give it, not realizing the position he is putting himself in. He could have his license suspended or he could be fined. Don't put him in that position, and don't subject yourself and other passengers to a situation which could be hazardous to all on board, and perhaps to someone on the ground.

Remember, you may not cause any interference all across the country, but the ILS glide slope receiver is used only during the last few minutes of flight, and interference to these units may not be noticed until it is too late.

Jim Fisk, W1DTY
editor

75 cents

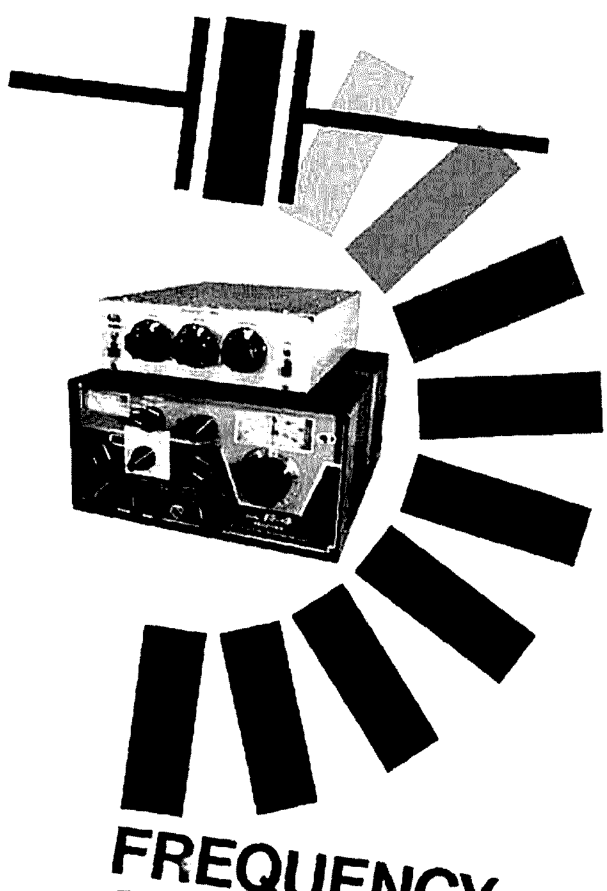
hr

focus
on
communications
technology . . .

ham radio magazine

AUGUST 1972

this month



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for the Drake R-4 receiver

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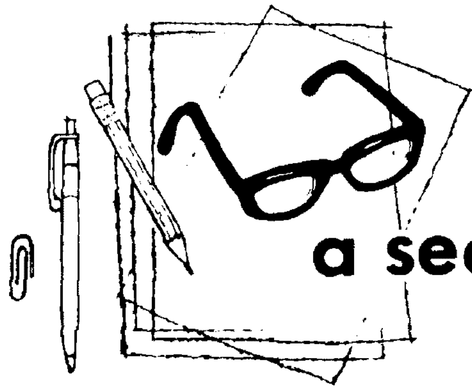


a second look

by jim fisk

The newest solid-state device to come down the electronics turnpike offers some interesting circuit possibilities with existing vacuum-tube equipment. Although the new Teledyne junction-fet *Fetron* is not a new device so much as it is a new application of old principles, it offers a simple solid-state plug-in replacement for many vacuum tubes.

At the present time there are only two types of Fetrons available, a triode, the TS12AT7, and a pentode, the TS6AK5. These Fetrons are designed to replace, respectively, the 12AT7 and 6AK5. Both of these tubes are members of large families of similar types, so the two available Fetrons are usable in many different circuits.

The secret to Fetron operation lies in the development of high-voltage junction fets which will operate without breaking down with the high-voltage power supplies found in most vacuum-tube equipment. Most Fetron circuits are built with two fets in a cascode arrangement to obtain high gain; the basic circuit was described earlier in *ham radio.* *

The Miller-effect capacitance is minimized in the Fetron by using an ultra-low-capacitance, high-gain transistor such as the 2N3823 in the input, with a high-voltage fet such as the 2N4881 at the output. The complete circuit goes into a metal can that has the same pin configuration as the tube it replaces.

Although the current cost of Fetrons is relatively high ($15.60 for the TS12AT7 and $12.50 for the TS6AK5), they make good sense as replacements for tubes in communications equipment because their characteristics don't drift with use, making periodic adjustment and alignment unnecessary, and they have extremely long life. In fact, the estimated life for a Fetron has been calculated at 30 million hours – that is equivalent to running them 24 hours a day, 365 days a year, for over 3400 years!

In addition to higher amplification factors, Fetrons feature noise figures much lower than many equivalent tubes, and much lower heat dissipation; lower heat dissipation because the Fetron requires no filament supply. The lower heat characteristic should be particularly useful in reducing drift in your vfo.

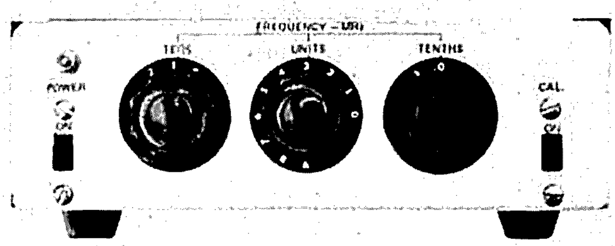
Although most Fetrons are designed for the lower frequency ranges, Fetron pentodes have been operated as high as 500 MHz. Charlie Spitz, W4API, has been experimenting with Fetrons in amateur communications equipment, but he has been hampered somewhat by the limited number of different Fetrons. However, this will be solved in the near future.

Teledyne has already built Fetron equivalents of the 6JC6 and 6EW6, and these can be combined with derivatives of the TS12AT7 and TS6AK5 to provide plug-in replacements for a great variety of tubes.

Next on the list of Fetron replacements will be power pentodes such as the 6AQ5 and 6V6, as well as remote-cutoff pentodes such as the 6BA6. With high-volume production and simplified packaging, Fetrons could well become low-cost replacements for most vacuum tubes. Watch the pages of *ham radio magazine* for the latest developments with this interesting new device.

Jim Fisk, W1DTY
editor

*W1DTY, "ham notebook," *ham radio*, February, 1970, page 70.



frequency synthesizer

for the Drake R-4 receiver

This frequency synthesizer converts the Drake R-4 series of amateur-band receivers to general coverage receivers that tune from 1.5 to 30 MHz

Robert S. Stein, W6NBI, 1849 Middleton Avenue, Los Altos, California 94022

The vast majority of receivers in use in ham shacks today cover the amateur bands only, with the possible addition of one of the WWV frequencies. While this may be entirely adequate for normal operation, there are times when a general-coverage receiver is a very useful adjunct. Being able to receive WWV on only one frequency is often unsatisfactory, since obviously the selection of the optimum receiving frequency depends on the time of day and the distance from Fort Collins.

There are many other uses to which a general-coverage receiver can be put. Tracking down harmonics and other spurious frequencies can be simplified if you have such a receiver. Calibration of some home-built test equipment may require a receiver which covers frequencies other than the ham bands. And it is often desirable to use a receiver as a tunable i-f with vhf converters whose output frequencies are *not* in one of the high-frequency amateur bands.

While reviewing some of the literature which has been published describing the newer integrated circuits, it occurred to me that a frequency synthesizer might be a practical method for converting my Drake R-4 to general coverage without a separate crystal for each 500-kHz range. Frequency synthesis techniques have

been around for some time, involving circuitry which has been complex and difficult. However, recent developments in integrated circuits have simplified the techniques to a point whereby the frequency synthesizer is a practical construction project for the relatively experienced amateur.

Before proceeding with the description of the frequency synthesizer which makes all-band coverage possible without modifying the receiver, a review of the receiver conversion process is in order. Although there are differences in the actual circuits of the Drake R-4, R-4A, and R-4B receivers, the conversion frequencies are identical; therefore, this discussion is applicable to any one of the three receivers.

If we disregard the pre-mixing arrangement used by Drake, the high-frequency oscillator requirements are simply that the oscillator be tuned 11.1 MHz above the low-frequency end of the desired 500-kHz tuning range. Since the high-frequency oscillator is crystal controlled and untuned, this requirement is translated to a crystal of the proper frequency. For example, the 3.5- to 4.0-MHz band requires a 14.6-MHz crystal (3.5 plus 11.1 MHz), which is supplied with the receiver. Drake specifies that the receiver will tune from 1.5 to 30 MHz with the proper crystal, excluding 5 to 6 MHz which bridges the 5645 kHz first *intermediate* frequency. (The apparent discrepancy between this i-f and the 11.1 MHz difference between the incoming and oscillator frequencies is taken care of in the premixing scheme, and can be ignored for the purposes of this article.) Since the receiver covers this range in 500-kHz steps, because the vfo has a 500-kHz tuning range, it would require 58 crystals to cover the complete spectrum, starting with a 12.6-MHz crystal for the 1.5- to 2.0-MHz band, a 13.1-MHz crystal for the 2.0- to 2.5-MHz band, and finally ending with a 40.6-MHz crystal for the 29.5- to 30.0-MHz band.

A complete set of crystals would obviously be expensive. Furthermore, there would be the practical difficulty of using all the crystals at one time since the receiver has provisions for externally installing only ten crystals in addition to the five amateur-band crystals mounted internally.

However, if we can generate frequencies of 12.6, 13.1, 13.6, 14.1, etc. through 40.6 MHz, and substitute them for the crystal-controlled high-frequency

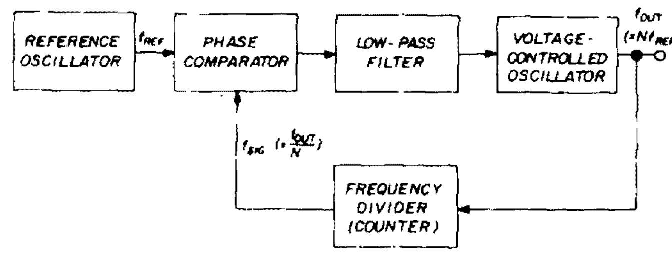


fig. 1. Basic phase-locked frequency synthesizer. The frequency divider is a variable-modulus, or programmable, counter.

oscillator, we can achieve all-band coverage within the specified range of the receiver. Thus, this becomes the requirement for the frequency synthesizer.

basic phase-locked loop frequency synthesizer

The integrated circuits to which I briefly referred are designed for use in phase-locked loops, one application of which is frequency synthesis. **Fig. 1** shows the basic phase-locked frequency synthesizer. A stable reference frequency is applied to one input of a phase comparator. The output of the phase comparator is a dc voltage, which passes through a low-pass filter and controls the frequency of a voltage-controlled oscillator (vco). This oscillator generates the desired frequency, which may be any multiple of the reference frequency. The vco output is also applied to a frequency divider whose function is to divide the vco output frequency to the same frequency as that of the reference oscillator.

Let's assume that the reference oscillator frequency is exactly 100 kHz and that an output frequency of 12,600 kHz is required. If we have a divider or programmable counter which will divide by 126, the signal input to the phase

comparator will also be 100 kHz when the vco output is exactly 12,600 kHz. This is accomplished by the phase comparator producing a d-c output which "tunes" the vco until it is exactly 12,600 kHz. The divided vco frequency is then exactly 100 kHz, the same as the reference frequency. Thereafter, the vco output will stay at 12,600 kHz; any variation from this frequency changes the signal input to the phase comparator, which in turn produces a dc output change and brings the vco back to 12,600 kHz. Thus, the output frequency is locked to the reference frequency, and has essentially the same stability as the reference oscillator.

By using a frequency divider which can be programmed, it is possible to obtain virtually any number of discrete frequencies which are integral multiples of the reference frequency, all of which are phase locked to the reference oscillator. The low-pass filter keeps the reference frequency from appearing at the vco and establishes the lock-up time of the loop. A rigorous analysis of the feedback loop and filter involves extensive mathematics and is beyond the scope of this article. For those of you interested in greater detail, references 1 through 3 provide a wealth of information and design data.

A block diagram of the R-4 high-frequency oscillator frequency synthesizer is shown in **fig. 2.** The loop reference is a 100-kHz crystal-controlled oscillator which uses a Signetics LU380A quad two-input NOR gate. The oscillator output is divided by ten (for reasons which will be discussed later) and the resultant 10-kHz signal is applied to the reference input of a Motorola MC4044P phase comparator. Because I used the 100-kHz calibrator crystal from the R-4 in the synthesizer, I provided a 100-kHz output jack on the synthesizer in order to introduce a calibration signal back into the receiver.

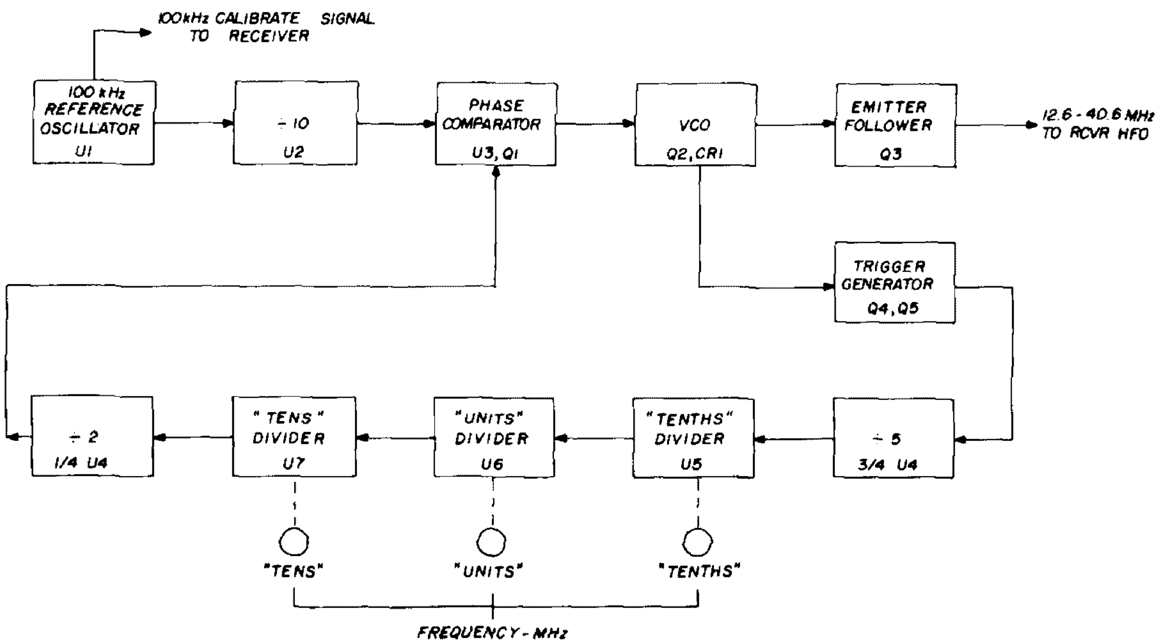
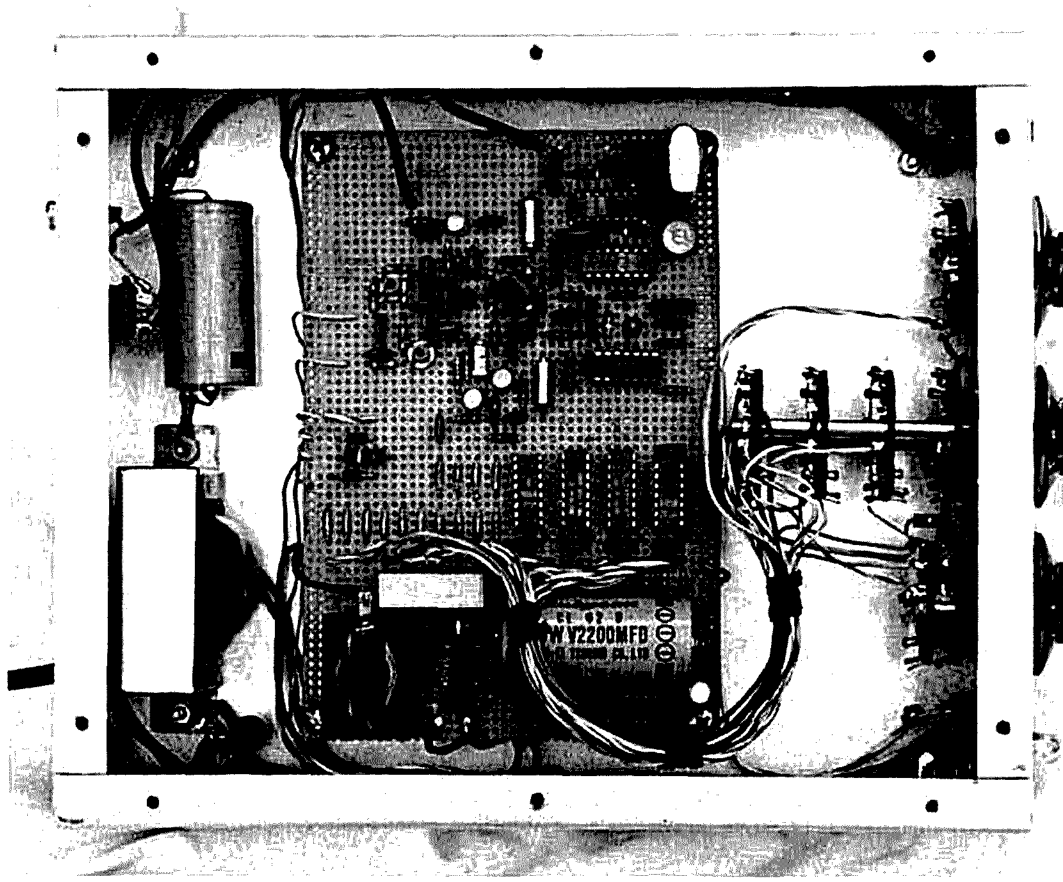


fig. 2. Block diagram of the R-4 hfo frequency synthesizer. The only tuning controls are the front-panel mounted tens, units and tenths rotary switches.

This is shown in **fig. 2**, and is covered in greater detail under the circuit description and interconnection portions of this article. It should be noted, however, that this technique will eliminate the 25-kHz calibration markers in the R-4B because the 100-kHz calibration signal is fed back into the receiver antenna input by way of the accessory socket on the receiver. If you wish to retain the 25-kHz markers, use a separate crystal in the synthesizer and delete the calibration circuit.

The low-pass filter, which appeared in fig. 1, is not shown separately in fig. 2 since part of it is integral to the MC4044P with the remainder made up of discrete components. The output of the phase comparator/filter is applied to a varactor diode which controls the frequency of the vco, a Fairchild SE3005 transistor. The vco output, 12.6 to 40.6 MHz in 500-kHz steps, is routed to the R-4 via a 2N709 emitter follower. The vco output also drives a trigger generator which uses a pair of 2N3478 transistors. The trigger generator is needed to convert the amplitude and waveform of the vco output to that required by the TTL integrated circuits in the frequency divider.

The frequency divider is a variable-modulus counter which can be programmed to divide by any factor between 1260 and 4060 in steps of 50. It is at this point that the reason for using a 10-kHz reference input to the phase comparator can be explained. If the reference signal were 100 kHz, the divider would need only a divide range of 126 to 406 in order to divide the vco output of 12.6 to 40.6 MHz down to 100 kHz. However, at the upper frequency limit of 40.6 MHz, the propagation delays through three synchronous counters arranged in a ripple-through configuration would prevent proper operation of the circuit.



Interior view showing construction of the synthesizer. The reference oscillator is at the top right of the board, the vco and trigger generator are at the upper center, and the frequency-divider ICs and steering diodes are to the left of the multi-section switch. Power supply components are mounted at the bottom of the board and on the chassis to the left of the board.

Therefore, the trigger generator output, which is at the vco frequency, is first divided by five in a quinary divider which consists of three of the four flip-flops in a Signetics 8290A high-speed decade counter. The remaining flip-flop is placed at the end of the divider chain, thereby dividing the vco output by a fixed factor of ten. Thus it can be seen that the reference signal must also be divided by ten if the reference and signal inputs to

the phase comparator are to be equal.

Although the trigger generator output could have been divided by ten before it was applied to the programmable counter, placing the binary divider at the end of the chain results in a square-wave output which is readily observable on oscilloscopes of limited frequency response and rise time.

Returning to the signal flow shown in fig. 2, let's assume that the desired vco output is to be 39.6 MHz, for a receiver tuning range of 28.5 to 29.0 MHz. The 39.6-MHz signal from the trigger generator is divided by five to a frequency of 7.92 MHz. To obtain a 10-kHz output from the divider chain, the dividers must be programmed to divide by 396. (Remember that there is a fixed divide-by-two at the end of the chain.) By programming the *tenths* divider to divide by six, the *units* divider to divide by nine, and the *tens* divider to divide by three, the output from the *tens* divider is 7.92 MHz divided by 396, or 20 kHz. This is divided by two in the binary section of the 8290A, resulting in a 10-kHz signal being applied to the phase comparator.

The designations *tenths, units* and *tens* used for the dividers refer to the panel markings of the rotary switches which are used to set the dividers. Each of the dividers is a decade counter which can be preset to count by any integer between one and nine. The *tenths* divider, being the first in the chain, determines the least significant figure in the total count. The *units* divider divides the output of the *tenths* divider; therefore its count (multiplied by 10) determines the middle figure in the three-digit count, while the count (multiplied by 100) of the *tens* divider determines the most significant figure in the total count.

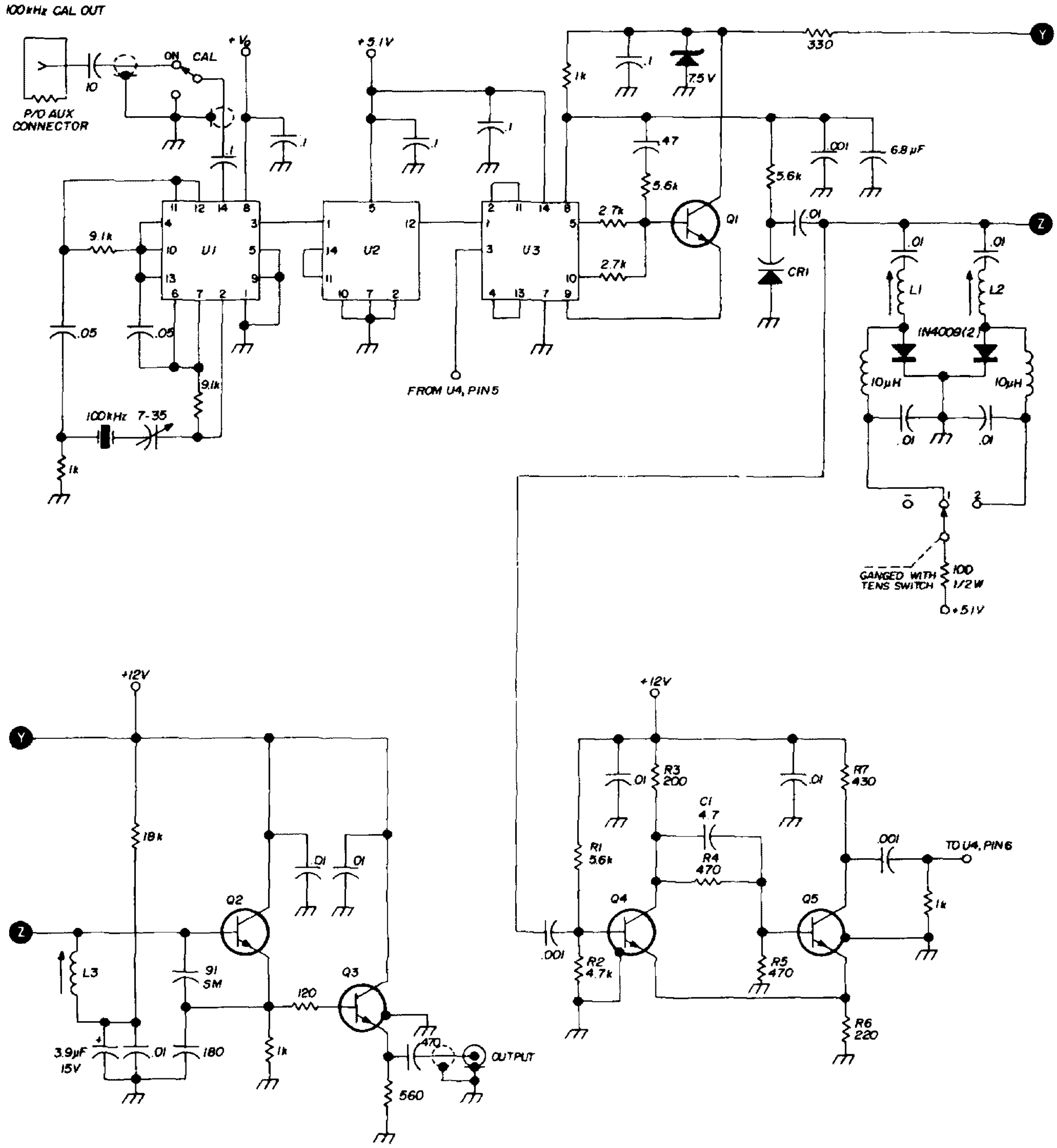
In order to obtain divide ratios of 1260 to 4060 in steps of 50, the variable-modulus counter must have a range of 126 to 406 in steps of 5 (since there is a fixed division by ten), i. e., 126, 131, 136, 141 . . . 401, 406. Therefore, the first, or *tenths,* divider is arranged to divide by one or six, as controlled by its associated switch. The *units* divider can divide by any number from one to nine, and the *tens* divider can be preset to divide by one, two, three or four.

Those of you who have looked at the front-panel knobs with a sharp eye will note that despite the foregoing explanation, the *tenths* switch is calibrated 0.5 and 0.0, while the *tens* switch is calibrated 2, 1 and ––. This was done so that the switches would indicate the receiver *tuning* frequency, not the hfo injection frequency. Obtaining this convenience required a considerable increase in switching complexity.

Correctly marking the *tenths* switch involved only applying the proper designations to the switch. When the *tenths* divider divides by one, the receiver tuning range starts at an integral megahertz; therefore that switch position is marked 0.0. When the divider divides by six, the receiver tuning range starts at a half-megahertz; thus that position is designated 0.5.

Explaining the *tens* switch is a bit more complicated. Remembering that the synthesizer output is always 11.1 MHz above the start of the receiver tuning range, we can see that the *units* divider will always divide by a factor of one greater than the *units* digit in the receiver tuning range. For example, if the receiver tuning is to start at *15.5* MHz, the *units* divider must divide by 6. This causes no problem except when the receiver tuning range starts at 9.0, 9.5, 19.0, 19.5, 29.0, or 29.5 MHz. The unit count of 9 means that the *units* divider must divide by 10, which is reserved for the *tens* divider. Therefore, the 9 position of the *units* divider switch actually causes the divider to divide by one and sets the *tens* divider to the next divide ratio.

The net result is that only three, not four, switch positions are needed for the *tens* divider. Even though the divider must divide by four (to obtain a division of *401* and *406* for 29.0 and 29.5 MHz respectively), setting the *units* switch to 9 sets the *tens* divider to the next count and eliminates the fourth switch position. The "––" position corresponds to a zero



CR1	varactor (Motorola MV1401)	Q3	2N709, 2N918, or 2N3563
L1	8 turns no. 28E, close wound on 5/32" diameter slug-tuned form	Q4,Q5	2N3478
L2	5 turns no. 28E, close wound on 5/32" diameter slug-tuned form	U1	quad 2-input NOR gate (Signetics LU380A or SE380A)
L3	8 turns no. 28E, close wound on 7/32" diameter slug-tuned form	U2	decade counter (Signetics N7490A, Texas Instruments SN7490N or Motorola MC7490P)
Q1	2N3391 or Motorola MPS6571	U3	phase-frequency detector (Motorola MC4044P or MC4344P)
Q2	Fairchild SE3005, 2N918, 2N3563, or 2N5179		

fig. 3. Schematic of the reference oscillator, phase comparator, vco and trigger generator circuits. Capacitors designated SM are silver micas and those with polarity indicated are electrolytic; all others are ceramic.

when the switches are set to a frequency of 9.5 MHz or lower.

circuit

Fig. 3 shows the frequency-generating circuits, the phase comparator and the trigger generator. The 100-kHz crystal oscillator uses two gate sections of U1, a quad two-input NOR gate; the remaining two sections are used as buffers. A logic diagram of the circuit is shown in fig. 4. Each of the two gates used in the oscillator is forced into linear operation by connecting a 9.1k resistor between the output and input; a 100-kHz signal is developed by connecting the crystal in series with the feedback loop from the output of the second gate to the input of the first. A 7-35 pF ceramic trimmer in series with the crystal permits adjustment of the crystal frequency. The remaining two gates buffer the output to the succeeding divide-by-ten stage and to the calibrate input of the R-4. The calibrate signal is disabled by grounding the output of the gate by means of the calibrate switch.

Returning to fig. 3, the 100-kHz signal is divided by ten in U2, and the resultant 10-kHz signal applied to pin 1 of phase comparator U3. This device consists of two separate functional parts, a phase detector and an amplifier, which must be connected externally. The outputs from the phase detector section appear at pins 5 and 10 and are applied to the loop filter consisting of two 2.7k resistors, a 5.6k resistor and a 0.47-μF capacitor. Transistor Q1 isolates the RC filter from the amplifier section of U3, the input of which is pin 9. The amplifier output from pin 8 is applied as the control voltage to varactor CR1. It was found necessary to add the 6.8 and 0.001 μF capacitors to reduce the amplitude of the 10-kHz reference signal which reached CR1. The supply voltage for the amplifier section of U3 is regulated at 7.5 volts by means of a zener diode and 330-ohm dropping resistor from the 12-volt supply.

The voltage-controlled oscillator consists of Q2, a high-frequency npn transistor, in a Colpitts circuit with varactor CR1 across the tank circuit. CR1 is a *Motorola MV1401 and has* a *ratio of* maximum-to-minimum capacitance of approximately ten, as compared to usual ratios of two to four for conventional varactors. (It also happens to be the most expensive single component in the entire synthesizer.) Despite the large capacitance ratio, the oscillator cannot cover the entire range of 12.6 to 40.6 MHz without switching. This is accomplished by diode switching, using a switch section ganged with the *tens* divider switch.

In the "——" position of the *tens* switch, the two 1N4009 diodes do not conduct, so coils L1 and L2 are each effectively in series with a 10-μH choke to ground. The high value of this inductance has only stray effect on the circuit; thus the oscillator frequency is determined by coil L3 and the tank-circuit capacitance. When the *tens* switch is set to position 1 or 2, one of the diodes is biased into forward conduction and

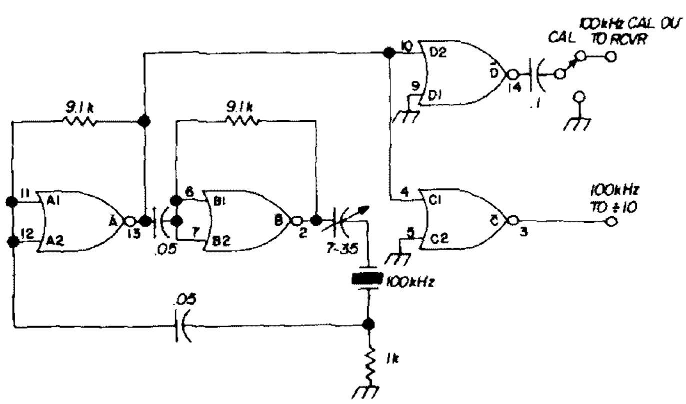
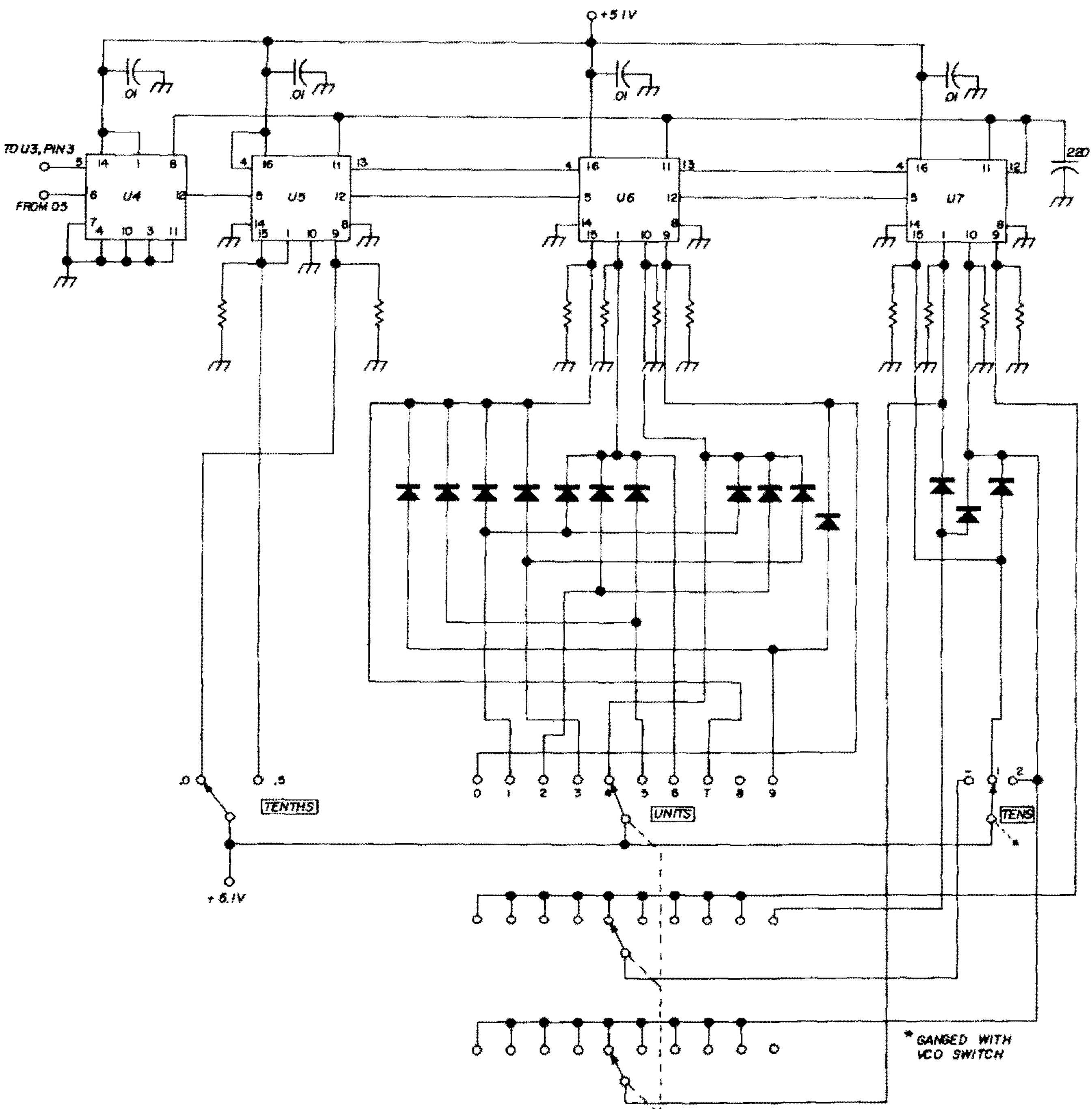


fig. 4. Logic diagram of the reference oscillator circuit.

brings the low end of the associated coil close to rf ground, shunting L3 and thereby lowering the tank-circuit inductance. The 100-ohm resistor in series with the switch arm limits diode current to a safe value.

The oscillator output is taken from the emitter of Q2 and coupled to the base of emitter follower Q3, which isolates the vco from the loading effects of the receiver and connecting cable.

Oscillator output is also taken from



U4	high-speed decade counter (Signetics 8290A)	U5,U6, U7	synchronous up/down counter (Signetics N74192A or Texas Instruments SN74192N)

fig. 5. Schematic of the frequency divider. All diodes are general-purpose germanium, all resistors are 910 or 1000 ohms, and all capacitors are ceramic. Switches are non-shorting rotary types.

the base of Q2 and coupled to the base of Q4, which with Q5, forms a Schmitt trigger. This circuit, when driven by the approximate sinusoidal output of the vco, generates a rectangular wave of several volts peak-to-peak amplitude with fast rise and decay times, which are required to toggle the TTL logic of the frequency divider. Since the Schmitt trigger is not common in amateur equipment, a brief explanation of its operation is in order.

In the absence of an input signal to the base of Q4, that transistor is normally cut off by the combination of base bias at the

junction of R1 and R2 plus the reverse bias at its emitter resulting from the collector current of Q5 flowing through R6. The resultant high positive voltage at the collector of Q4 forward biases Q5 to saturation. When an incoming signal of sufficient amplitude is applied to the base of Q4, the positive excursion drives the voltage at the collector of Q5 drops. Thus a rectangular output pulse is generated from a sinusoidal input, having rise and decay times sufficiently short to trigger the TTL decade counter which follows the Schmitt trigger.

As shown in fig. 5, the output of Q5 is applied to pin 6 of U4, which is the clock

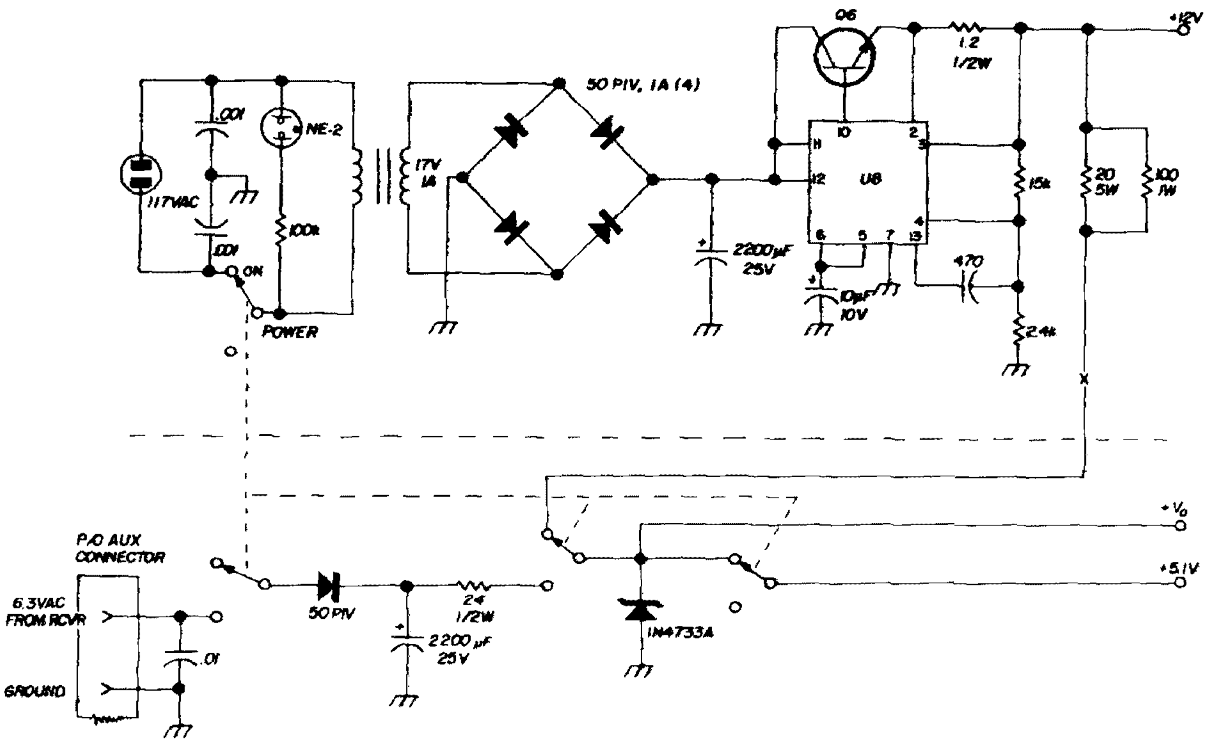


fig. 6. Schematic of the power supply. Power transistor Q6 is a 2N3054 or equal. U8 is a Signetics NE550A. See text for details of the power transformer and the meaning of point X.

base into conduction and causes collector current to flow. The drop in collector voltage is coupled via R4 and C1 to the base of Q5, reducing the base current and the collector current of Q5. The decrease in current lowers the voltage drop across R6, which results in increased collector current in Q4. This regenerative action causes Q4 to saturate quickly and cuts off Q5, producing an extremely fast rise time as the collector voltage of Q5 reaches the supply voltage level.

When the input signal becomes negative, the circuit operation reverses, cutting off Q4 and driving Q5 back into saturation. The regenerative action likewise results in a short decay time as the input for the quinary divider section of that decade divider. The output from pin 12 is applied to pin 5 of U5; U5, U6 and U7 are programmable decade counters set up as a ripple-through or variable-modulus counter. The count or divide ratio is determined by setting pins 1, 9, 10 and 13 of each IC high or low. This is accomplished by the rotary switches described previously; steering diodes are used to reduce the number of switch sections which would otherwise be required. The two extra sections of the *units* switch are needed to advance the count of the *tens* divider when the *units* switch is in position 9.

The output of the variable-modulus

counter is applied to pin 8 of U4, the clock input of the binary section of that device. This flip-flop generates a 10-kHz square wave from the 20-kHz input from the counters; the square wave is the signal input which is applied to pin 3 of U3 for phase comparison with the 10-kHz reference signal.

power supply

The power requirements of the frequency synthesizer are +12 volts at 65 mA and +5 volts (nominal) at 360 mA. The 12-volt supply must be well filtered to prevent frequency modulation of the vco. The schematic appears in **fig. 6.**

A full-wave bridge rectifier is supplied from a 17-volt, 1-ampere transformer. The positive output from the rectifier is fed to the input (pin 12) of voltage regulator U8. Since the current drain exceeds the current limitation of the regulator, power transistor Q6 is incorporated as a series-pass device. The output of the regulator is set to 12 volts by means of the 15k and 2.4k resistors connected to pins 3 and 4 of U8. The 5.1-volt supply for the logic elements is developed across a 1N4733A zener diode. Current limiting and overload or short-circuit protection is provided by monitoring the voltage drop across the 1.2-ohm resistor in series with the emitter of Q6.

As previously noted, the 100-kHz calibrator crystal from the R-4 was removed from the receiver and used as the reference oscillator, with provisions incorporated to reintroduce the 100-kHz signal back into the receiver. In order to enable the reference oscillator in the synthesizer for this calibrate signal when the synthesizer is not in use, an auxiliary power supply is included which keeps the reference oscillator energized whenever the receiver is on, whether or not the synthesizer is being used.

A 4-pole, double-throw slide switch is used as a power switch. In the on position, 117 volts ac is applied to the transformer, and all circuits in the synthesizer are energized. When the power switch is in the off position, a separate half-wave rectifier and filter are connected through a cable to the 6.3-volt pin of the accessory power receptacle on the rear of the R-4. When the receiver is on, this voltage is rectified and filtered, and applied to the same zener diode used in the main power supply of the synthesizer. However, only the reference oscillator circuit is powered from the auxiliary power supply.

If a separate crystal is used in the synthesizer and the calibrator crystal is left in the receiver, the auxiliary power supply and its switching circuit can be eliminated. This is accomplished by deleting all of the circuit below the dashed line in **fig. 6** and connecting the 1N4773A zener diode between point X and ground. The 5.1-volt supply is obtained from point X. Pin 8 of U1 (see **fig. 3**) should be connected directly to the 5.1-volt supply.

construction

Construction of the frequency synthesizer is shown in the various photographs. An 8 x 10 x 2½ inch aluminum chassis is inverted and used as a base. Four rubber bumper feet mounted on the bottom prevent scratches from the screws and nuts which fasten the parts to the chassis.

The three rotary switches, the power and calibrate slide switches and the NE-2 pilot lamp are mounted on the front apron. On the rear apron are the coax output connector and a three-pin miniature connector used for the 6.3-volt and 100-kHz calibrate signal connections between the synthesizer and the receiver. The power transformer and the filter capacitor for the auxiliary power supply are mounted directly on the chassis. All other parts are mounted on a piece of perforated Vector board, copper-clad on one side. The board measures 4½ x 6½ inches overall and the hole spacing is 0.1 inch. This spacing is important, since it matches the pin spacing on DIP integrated circuits. The use of copper-clad board provides a ground plane for the rf and high-speed logic circuits.

The ungrounded leads of the various

components are soldered to insulated pads cut into the board by means of a Vector P-138C pad-cutting tool, and then connected by conventional wiring. The usual precautions of short, direct leads should be observed; although the highest radio frequency involved *is* only 40.6 MHz, vhf techniques should be followed because of the high-speed logic pulses. The integrated circuits may be soldered directly to the board, or sockets or Molex pins may be used. My personal preference is the latter, since they are inexpensive and eliminate the very difficult job of unsoldering an IC should it be necessary.

The power supply series-pass transistor is mounted in a Thermalloy 6166-B heat sink, which is then mounted on the perf board. Although the heat sink is quite small, it is adequate for the power dissipated in the transistor and holds the case temperature to about 60° C, well within the temperature-power dissipation curve limits.

All connecting leads to and from the board should be soldered in place on the board before connecting the other ends. Leave plenty of wire on each lead to make the connections to the parts which are *not* on the board so that if any troubleshooting must be done it will not be necessary to unsolder the wires. Wiring for the output and calibrate signals is made with RG-174/U coax and should be kept reasonably short. All other wiring carries dc *only*, making lead length uncritical. The board is mounted on four metal standoff posts, which ground the copper side to the chassis.

The power transformer should supply 16 to 17 volts ac at approximately 0.5 ampere. Any higher voltage only results in greater heat dissipation in the series-pass transistor. The simplest way to obtain a transformer is to modify one of the inexpensive 24-volt, 1-ampere transformers which are readily available. First connect the primary of the unmodified transformer and measure the secondary voltage. Then disconnect the transformer and carefully cut through the outer paper insulation to expose the outer winding layer of the secondary. Remove ten turns from this winding and remeasure the secondary voltage. The difference between the new lower voltage and the original secondary voltage, divided by ten, is the volts per turn of the secondary. From this figure calculate the number of turns required for a voltage equal to the difference between the original secondary voltage and the desired 17 volts, and remove this number of turns from the secondary. After checking the new secondary voltage, wrap the transformer windings with insulating tape.

There is a short-cut to this procedure if you can see and count the *number of layers* comprising the secondary. This is generally possible by closely examining and probing the open ends of the windings. Assume that you have determined that the secondary has eight layers, which means that each layer develops approximately 3 volts (24 divided by 8). Removing two and a half layers should result in a secondary voltage of 16.5 volts, which is just about right.

A cover for the chassis was made from aluminum cane-metal stock, which provides adequate ventilation for dissipating the heat generated. The numbered positions on the rotary switch knobs were made by using white number transfers on the black skirts of the knobs. Several heavy coats of Krylon fixative were sprayed on to keep the numbers from peeling off.

alignment and test

After all wiring and connections have been checked and rechecked, the synthesizer is ready for the few adjustments necessary to set the vco on frequency. The only test equipment absolutely necessary is a voltmeter, although a grid-dip oscillator and oscilloscope can be helpful.

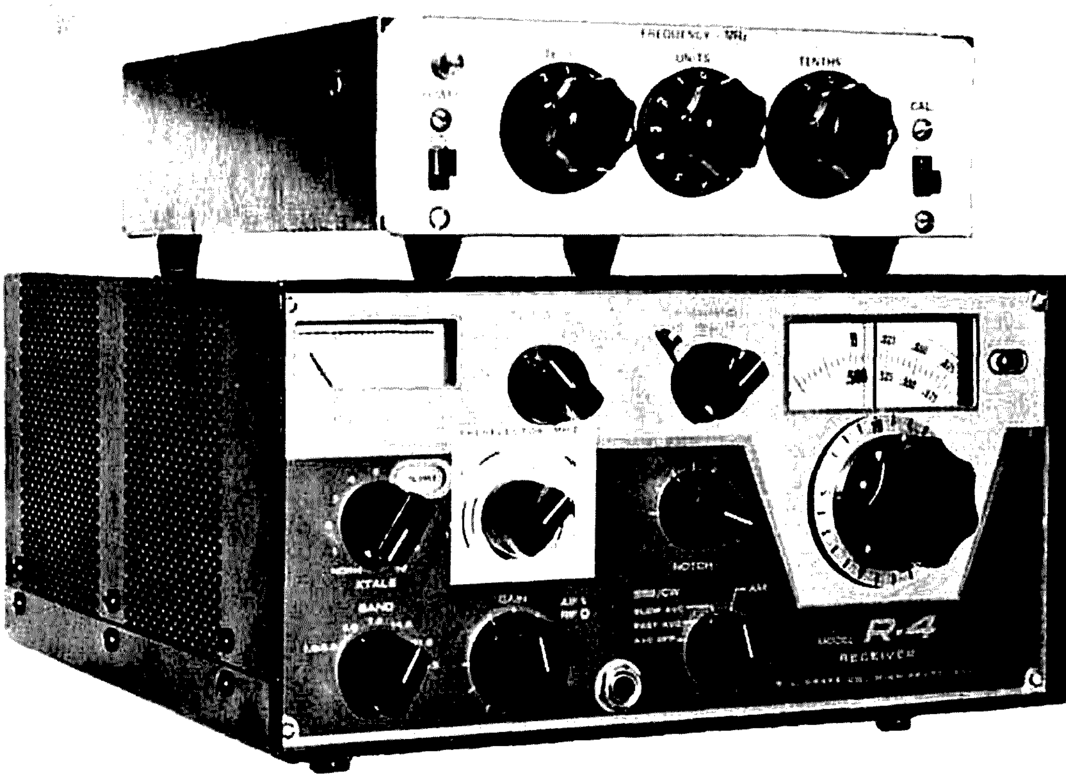
Apply power to the synthesizer and check the supply voltages to make sure that they are within five percent of the nominal values. Then make sure the reference oscillator is working by bringing a lead from the receiver antenna jack close to the crystal. Harmonics of the 100-kHz oscillator should be heard in the

receiver. The oscillator may also be checked with a scope; 100-kHz square waves should be observed at pin 3 of U1, and 10-kHz square waves should be present at pin 12 of U2. If the oscillator is not working, adjust the trimmer capacitor; it is not necessary for the crystal to be oscillating at exactly 100 kHz at this time.

Connect a voltmeter between pin 8 of U3 and ground. Set the rotary switches for a frequency of 9.5 MHz and move the tuning slug of L3 into the coil until a maximum positive voltage is obtained. This should be 7.5 volts, the same as the zener-regulated voltage applied to U3 and Q1. Slowly tune the slug out of L3 until the meter reading drops to 7.1 volts, or about 0.4 volt less than maximum.

Rotate the *units* switch toward zero, noting that the voltmeter reading drops with each switch position until position 1 is reached. (If the voltage does not drop between switch positions 1 and 0, it is nothing to be concerned about. The vco may not tune this low, but the switch position corresponds to a receiver tuning range starting at 500 kHz, which is outside the receiver preselector tuning range.) If you have a grid dipper available, couple it to L3. It should indicate 12.6 MHz in position 1 of the *units* switch, and 20.6 MHz in position 9.

Next, set the rotary switches for a frequency of 19.5 MHz, and adjust the slug in L1 for a voltmeter reading of approximately 6.5 volts. *Do not touch the slug in L3.* Again turn the *units* switch toward zero and note that the voltage drops with each switch position, including position 0 this time. Turn the *tenths* switch to its 0.0 position and make sure that the voltage also decreases at this step. A grid dipper coupled to L1 should



The frequency synthesizer atop the Drake R-4 receiver, which has the new general-coverage preselector scale on its front panel. Operating the equipment in this physical configuration is not recommended because the synthesizer interferes with convection flow from the R-4.

indicate 21.1 and 30.6 MHz for the minimum and maximum *units* and *tenths* switch settings.

Finally, set the rotary switches for a frequency of 29.5 MHz and repeat the preceding adjustment for L2. The frequency range of the vco with the *tens* switch in position 2 is 31.1 to 40.6 MHz.

As a further check, a scope can be connected to pin 3 of U3. If the scope has a triggered sweep, the signal at pin 12 of U2 should be connected to the external trigger input of the scope. A 10-kHz square wave should be displayed on the scope for each and every position of the frequency selection switches on the synthesizer. This completes the entire alignment procedure, except for setting the crystal to exactly 100-kHz after the synthesizer has been connected to the receiver.

operation

The vco output from the synthesizer is fed into one of the auxiliary crystal sockets on the rear of the receiver, so that no modification of the receiver is required. The auxiliary crystal sockets are arranged in two rows of five each, one above the other. The socket holes in these two rows which are adjacent to each other (the bottom holes in the top row of sockets and the top holes in the bottom row of sockets) will be referred to as the *inside* holes.

If the synthesizer is to be used with a Drake R-4, build a cable as shown in fig. 7A, and insert the crystal-base end into one of the auxiliary crystal sockets so that the pin connected to the coax shield is in the *inside* hole of the crystal socket. The 150-ohm resistor keeps the receiver hfo from oscillating on its own.

If the synthesizer is to be used with the Drake R-4A or R-4B, use the cable shown in fig. 7B. The pin on the center conductor goes to the *inside* hole of one of the auxiliary crystal sockets, while the coax shield is grounded to the receiver chassis. The ground stud on the receiver is a convenient terminal for the shield.

Fig. 7C shows the cable to be used when the auxiliary power supply and 100-kHz calibrate circuit have been included in the synthesizer. The lead lengths are not critical, but there is little point in having the cable appreciably longer than the output cable.

After the receiver and synthesizer have been interconnected, turn on both and

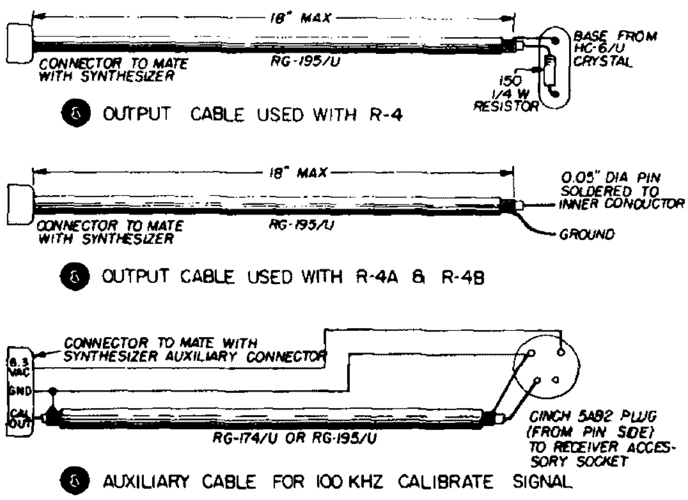


fig. 7. Interconnecting cables.

set the synthesizer switches to one of the WWV frequencies suitable for good reception. Set the receiver *xtals* switch to the position which corresponds to the crystal position now occupied by the synthesizer output cable. Refer to the preselector tuning chart in the receiver manual and set the receiver *band* switch and *preselector* tuning to the settings specified for the frequency selected. WWV should be heard when the receiver vfo is tuned to zero. Then allow the synthesizer and receiver to warm up for about a half-hour.

If you have incorporated the calibrate function in the synthesizer, turn on the calibrate switch and carefully adjust the 100-kHz crystal trimmer capacitor in the synthesizer to zero-beat the calibrate signal with WWV. It may be necessary to retune the receiver slightly when doing this because changing the frequency of the crystal changes the synthesizer output frequency which is now the receiver hfo frequency.

On the other hand, if you have retained the calibration function in the receiver, it will be necessary to bring a lead from the receiver antenna (which is available from the accessory socket on the rear apron of the receiver chassis) close to the 100-kHz reference oscillator in the synthesizer. Then adjust the crystal trimmer as described in the preceding paragraph.

That completes the installation and you now have a general-coverage receiver. Simply set the synthesizer switches to the frequency you want, set the receiver *band* switch and *preselector* control to the appropriate positions, and tune. As a final touch, add a new *preselector* scale so that you don't have to consult or memorize the preselector chart in the manual. Fig. 8 is a full-scale reproduction of the new *preselector scale*, and may be cut *out* or photocopied if you prefer not to mutilate the magazine.

Use a standard ¼ inch paper punch to cut a hole for the preselector shaft. Remove the *preselector* knob and attach the scale to the front panel of the receiver, using either a spot of rubber cement or a small piece of double-sided sticky-back tape in each corner. The scale shows the approximate setting of the *preselector* control; the large number at the end of each scale segment indicates the *band* switch position to be used.

When used with the R-4, the synthesizer may be left on whether it or one of the receiver crystals is generating the hfo frequency. When the synthesizer is used with the R-4A or R-4B, it must be turned off if the receiver *xtals* switch is not in the synthesizer position. The synthesizer and a crystal are both connected to the hfo in this case, so that possible spurious frequencies may result if the synthesizer is left on.

conclusion

This frequency synthesizer has been in operation for several months and fulfills the requirement of converting my R-4 receiver to general coverage. It is, however, a first attempt at synthesis techniques and can undoubtedly be improved. Despite considerable experimenting with the loop filter, there remains some frequency modulation of the vco at the reference oscillator frequency. This manifests itself as a weak spurious signal 10-kHz either side of true frequency, but is generally not noticeable since it appears

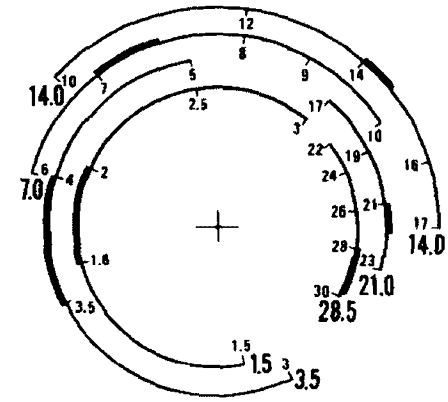


fig. 8. Receiver preselector scale.

to be 50 to 60 dB down. It is entirely possible that this modulation is enhanced by stray coupling, since all circuits are located on one board without any shielding whatsoever. Enclosing the vco in a shielded compartment might further attenuate the 10-kHz signal reaching it.

Digital circuitry is becoming more and more a fact of life for the amateur experimenter. A project such as this, involving digital and rf circuits, provides a welcome change from the usual, and yet is not completely alien to those without digital experience. It is offered, not so much as a "how to" project, but as a challenge to improve on it or adapt some of the ideas to other purposes.

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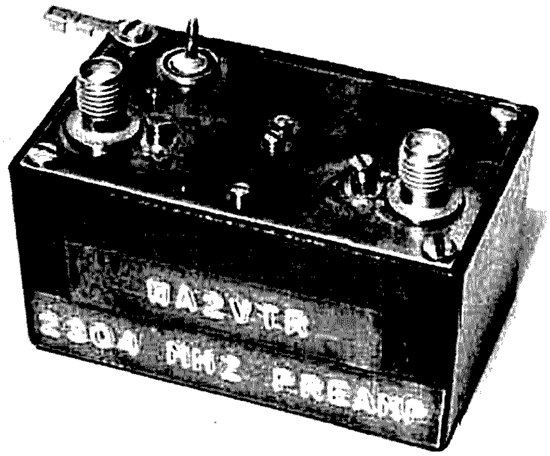


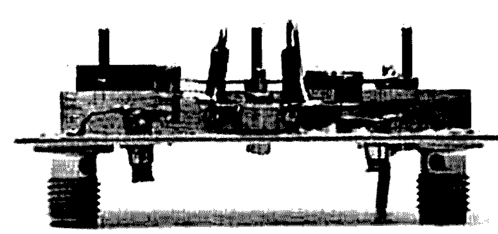
solid-state 2304-MHz preamplifier

High-performance preamplifier for 2304 MHz provides 8 dB gain and 5.7 dB noise figure

Dolph Vilardi, WA2VTR, 14 Oakwood Terrace, Spring Valley, New York

Receivers and converters for the uhf range have traditionally used mixer-type front ends because of the unavailability of devices which could provide useful amplification and sufficiently low noise figure. The noise figures provided by uhf mixers are limited not only by the devices available, but to a very large extent by the circuit in which they are used. For simplicity, amateurs usually use single-ended diode circuits.

Until recently, if uhf signal amplification was essential, parametric amplifiers were about the only way to obtain a



Side view shows vertical mounting of resistors. If 1/8-watt resistors are available, these components can be smaller. The partition mounting screws are longer than necessary. The miniature potentiometer is mounted with a drop of epoxy. It is adjusted through a hole in the side of the Pomona box.

substantial improvement in performance. However, several years ago transistors became available which had useful gain and reasonable noise figures on 1296 MHz. The cost of these early uhf transistors was high, but they eventually found their way into many high-performance 1296-MHz converters. Now the price of these devices is substantially lower, and the day of the parametric amplifier on 1296 and 2304 MHz, with its attendant isolators and circulators, is gone.

The success of the Nippon Electronics 766A as an 1296-MHz rf amplifier,[1] and its published performance specifications for 2000 MHz made it an interesting candidate for 2304 MHz. The proposed 2304-MHz pre-amplifier had to be easy to build, so it could be duplicated by other amateurs, and it had to offer improved performance over the straight diode mixer front end. With this in mind, I first tried to modify the pi-network circuit used in the 1296-MHz preamplifier. However, all attempts to develop a 2304-MHz preamplifier using this concept were unsuccessful, primarily because of problems with self oscillation.

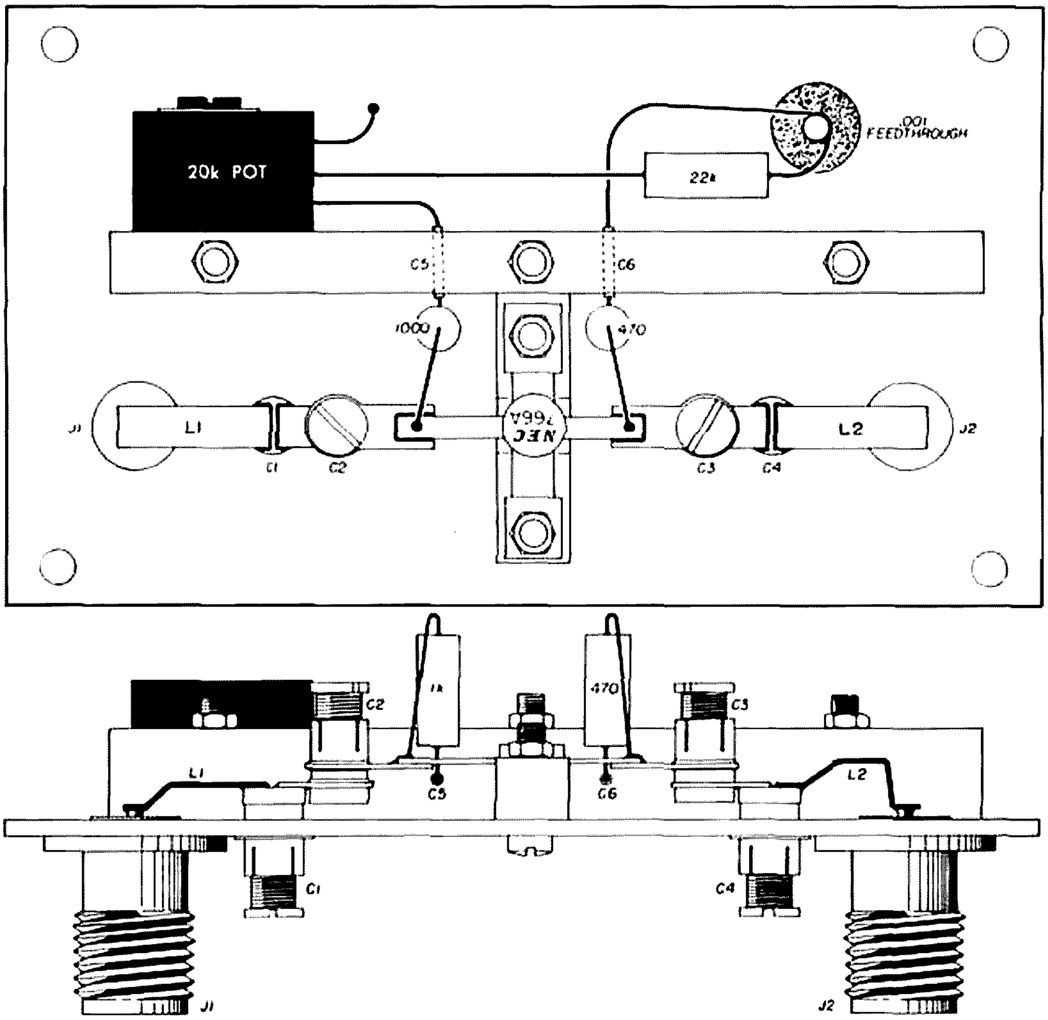
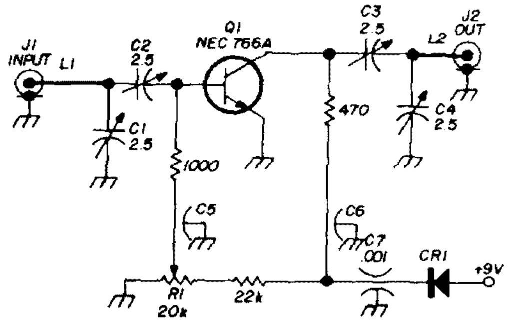


fig. 1. Layout of the 2304-MHz preamp. This drawing is twice actual size. Feed-through capacitors C5 and C6 consist of Teflon-covered wire, fit snugly into hole in 1/8" chassis partition.

About this time, my attention was called to a capacitor being marketed by the Johanson Corporation which was designed specifically for uhf impedance-matching purposes. This capacitor, the Johanson model 4991, is a miniature dual-variable capacitor with two 0.8- to 10-pF sections.

After conversations with Mr. Harvey Bruning of the Johanson Corporation



C1,C2	0.5-2.5 pF piston trimmers
C3,C4	Johanson Gigatrims JMC 6927
C5,C6	gimmick feedthrough capacitors (see fig. 2)
C7	0.001 μF feedthrough
R1	miniature 20,000-ohm potentiometer
L1	brass strip, 0.001" thick, 3/32" wide, 5/16" long
L2	brass strip, 0.001" thick, 3/32" wide, 3/8" long, formed as shown in fig. 2
J1,J2	miniature coaxial connectors (type SMA)

fig. 2. Schematic diagram of the 2304-MHz preamplifier. Transistor Q1 is a Nippon Electronics type V766A.

regarding a T-network arrangement using the new dual capacitors, he suggested a new device marketed by their company called the *Gigatrim.* These are very small concentric air-type capacitors which lend themselves to the circuit I wanted to use for the 2304-MHz preamplifier.

Mr. Bruning sent along some *Gigatrims* for experimental use, and with a layout suggested by him, the preamplifier seemed to work quite well. With the *Gigatrims* I was able to keep the overall size small enough so lead inductance did not become a limiting factor. Tuning the preamplifier is quite easy, and when built as described in this article, it shows absolutely no signs of instability.

Construction of the preamplifier should be obvious from the drawings and photographs, and should present no difficulties to the experienced uhf worker. Although Johanson states that the *solder* used in their products will withstand relatively high temperatures, I used TIX high-conductivity, low-temperature solder which melts at 250° and flows easily.* When using this type of solder, which contains indium, you must be careful because the liquid solder flows readily, and can get into the threads of the capacitor where it doesn't belong. According to Mr. Bruning, regular 60/40 solder *should work* well.

When you are building the 2304-MHz preamplifier, it is important that you follow directions closely, particularly in regard to the dimensions of the holes in the center partition which make up the feedthrough capacitors (C5 and C6). Close fit is essential for these capacitors to work as required.

Almost any miniature potentiometer may be used for R1. The *Cubic* potentiometer shown in the photographs is very compact and is easily mounted with a drop of epoxy. For the fixed resistors, use the smallest size that is available, 1/8 watt if possible.

tuning

The preamplifier is initially tuned up by injecting a 2304-MHz signal through a suitable attenuator and filter, and tuning all capacitors and adjusting the bias control (R1) for maximum gain. This technique is adequate because the noise figure

*TIX low-temperature solder is available from the Brookstone Company, 5 Brookstone Building, Peterborough, New Hampshire 03458.

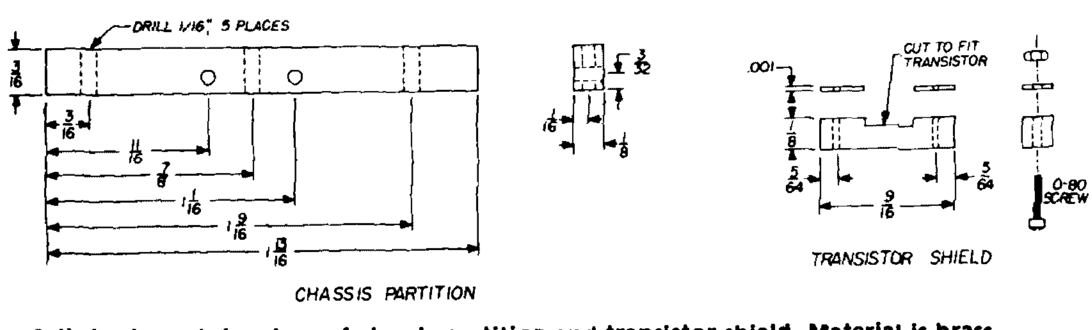


fig. 3. Full-size layout drawings of chassis partition and transistor shield. Material is brass.

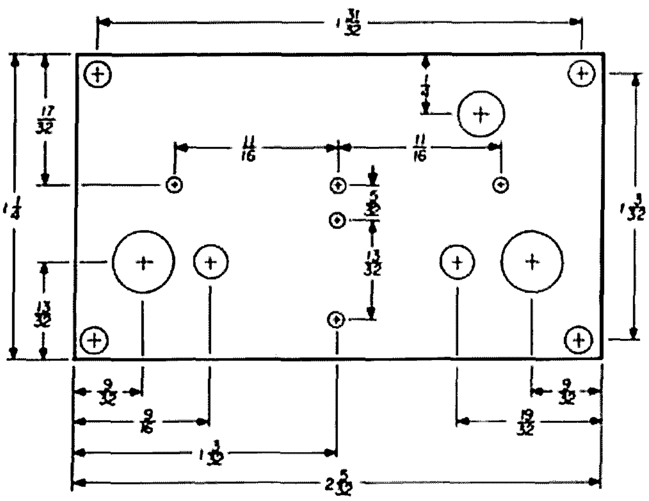


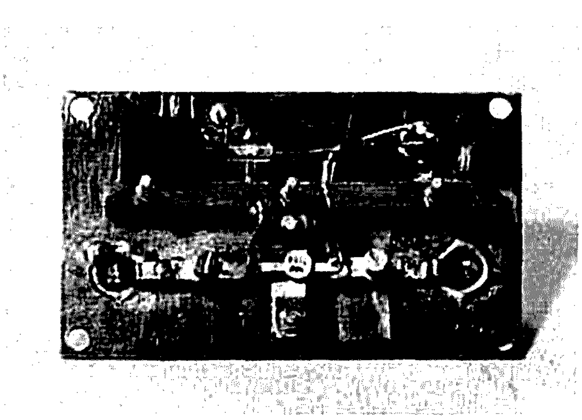
fig. 4. Full-size drill template for the chassis.

of the converter with which it will be used is probably higher than the gain of this single-stage preamplifier.

A simple, low-level signal source suitable for tuning up the 2304-MHz preamplifier is shown in fig. 5. This unit is quite similar to the design used with the 1296-MHz preamplifier, except that a 96-MHz overtone crystal is used in the oscillator. Also, it's desirable to feed the output of the multiplier through another resonant trough which serves as a signal filter. This is shown in fig. 5.

Final tuning should be accomplished with the preamplifier in the system, with the antenna connected. The low-level 2304-MHz signal is then fed in through the antenna.

I obtained a noise figure of 5.7 dB with this circuit, but there is little doubt that a selected transistor can improve this by at least 1 dB. Gain of 7 to 8 dB is



This photo shows the thickness of the partition which is important because it determines the capacitance of the feed-through capacitors, C5 and C6.

about all that can be expected from this transistor at 2304 MHz.

However, Nippon Electronics supplied a type V578A. This device offers a tremendous improvement at 2304 MHz in the same preamplifier. Noise figure measured at 3.8 dB with the V578A, as opposed to 5.7 dB with the V766A – all with no changes except tweaking. At 1296 MHz, the noise figure was 2.3 dB for the V578A vs 2.8 dB for the V766A.

Soon after the preamplifier was built, I wrote to Fairchild and asked if they would be interested in supplying an MT4578 transistor which, from the published specifications, looked like it

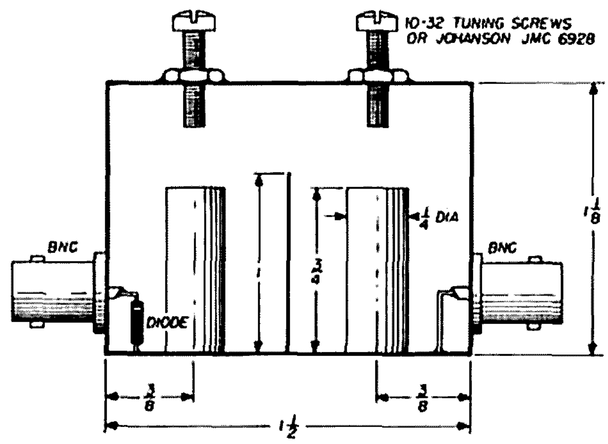


fig. 5. Weak-signal source for tuning up the 2304-MHz preamplifier. Small varactor diode or 1N914 is useful as multiplier, but try anything! Although 10-32 screws can be used for tuning screws, the Johanson JMC 6928 capacitive tuning screw is designed specifically for this purpose, and is virtually noise free. Cavity is 1" deep.

would give improved performance. After waiting some time, I have not heard anything from Fairchild, so I haven't made any experiments with the MT4578.

I wish to express my appreciation to the Johanson Corporation and Nippon Electronics for their cooperation, and particularly to Mr. Harvey Bruning of Johanson whose suggestions were quite helpful.

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2. W. Stanton, K2JNG, D. Moser, WA2LTM, D. Vilardi, WA2VTR, "Solid-State 2304-MHz Converter," *ham radio*, March, 1972, page 16.

ham radio

inexpensive audio filters

Surprisingly effective audio filters are possible using non-critical readily available components

Bill Wildenhin, W8YFB, 41230 Butternut Ridge, Elyria, Ohio 44038

Ed Noll's excellent summary of direct conversion receivers in the November, 1971 *ham radio*, correctly points out the common failing of inadequate audio selectivity.* In the course of developing some simple, inexpensive equipment for novices, I have found it possible to obtain *really excellent selectivity* without resorting to rigorous filter design and without careful selection and matching of parts. A sufficient number of these filters, and variations of them, have been made to insure reproducibility.

basic circuit

Fig. 1 shows the basic circuit — just three constant-K pi *sections in cascade*. For those who are not familiar with filters, this filter is a low pass filter — it

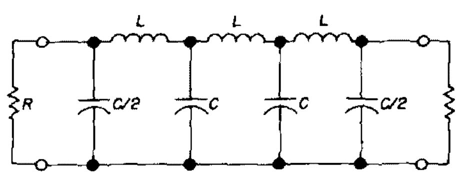


fig. 1. The basic circuit of a three section, constant-K pi-section low-pass filter. Some typical practical values for filters like this are given in table 1.

passes all frequencies from zero to some desired cutoff frequency. The impedance into and out of the filter is fixed by the chosen values of inductances and capacitances, and in this case the impedance in and out is the same value. **Table 1** was made up to allow choice of easily available capacitors. In all cases the inductors are surplus 88-mH toroidal units. **Fig. 2** shows how the leads are connected for use. As supplied, all four leads are usually brought out individually.

The curves in **figs. 5** and **6** are indicative of the results to be expected with off the shelf components. Some filters were built with used tubular paper capacitors pulled from scrap TV sets. Others used new components. If you purchase new components, the Cornell-Dubilier 100 V Mylar capacitors in their inexpensive WMF series are compact and reliable. *In no case should you use electrolytics.*

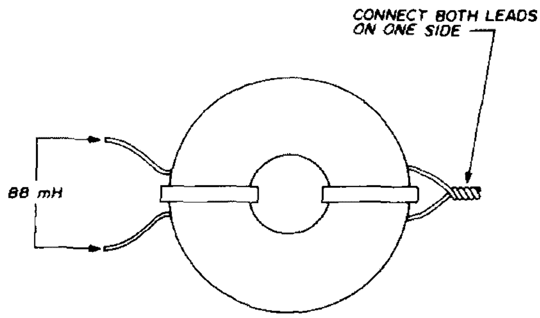


fig. 2. Proper way to connect leads to surplus 88-mH toroids for use in simple audio filters.

Fig. 3 shows a variation that has been successfully duplicated by several fellows.

table 1. Component values for low pass filters with different cutoff frequencies. Component values have been rounded off, but are accurate enough to produce excellent results in the circuit of fig. 1.

cutoff freq. Hz	R ohms	C μF	C/2 μF
550	150	4.0	2.0
800	200	2.0	1.0
1000	300	1.0	0.5
2300	650	0.2	0.1
2750	750	0.15	0.075

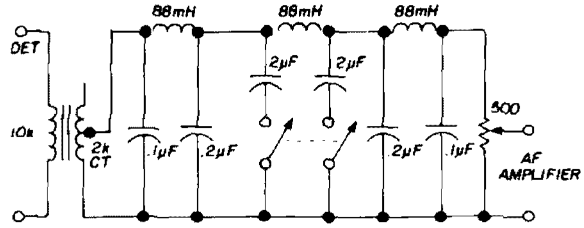


fig. 3. Audio filter whose cutoff frequency can be switched between two values.

This filter allows a higher cutoff frequency for phone reception and a low frequency for CW work. A double-pole single-throw slide switch is adequate for this application. The interstage transformer is a low cost import from Radio Shack. The detectors used either a Motorola MC1550 or RCA CA3028 integrated circuit. The 500-ohm volume control terminated the output of the filter and feeds a solid-state amplifier described in *ham radio,* March 1970.[2]

construction

Fig. 4 shows a compact suggested layout. For those who have no etching facilities, use Instant etch. No chemicals, no mess. Draw the layout on the copper, and score the copper with a sharp hobby knife. Carefully lift a corner of the

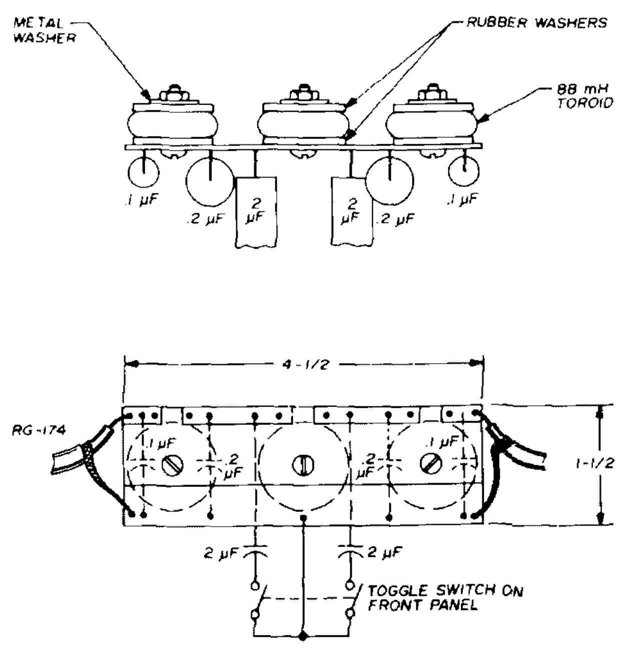
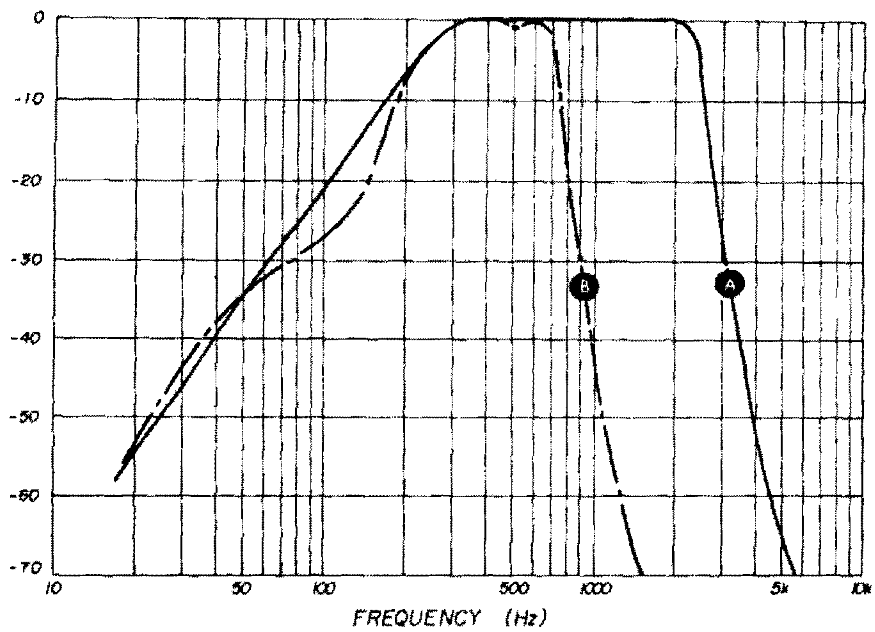


fig. 4. Suggested compact layout for the audio filter of fig. 3.



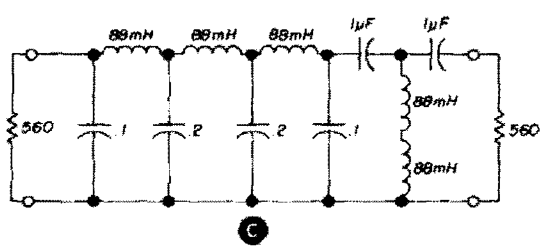


fig. 5. Performance curve of the low-pass filter shown in 5C. Curve A is of a filter similar to 5C, but with all 0.1-mF capacitors increased to 1.0 mF, and all 0.2-mF capacitors increased to 2.0 mF.

undesired copper with a knife edge, and then peel it off with a pair of long-nose pliers. If you wind the copper up as if opening a sardine can the work goes easier. You can also use thin, unplated board material and substitute pieces of number 22 tinned solid wire for the various busses. When mounting the toroids, it is best to use flexible washers. These can be cut from inner tube rubber, Teflon or soft vinyl plastic. The mounting screw should be covered to prevent

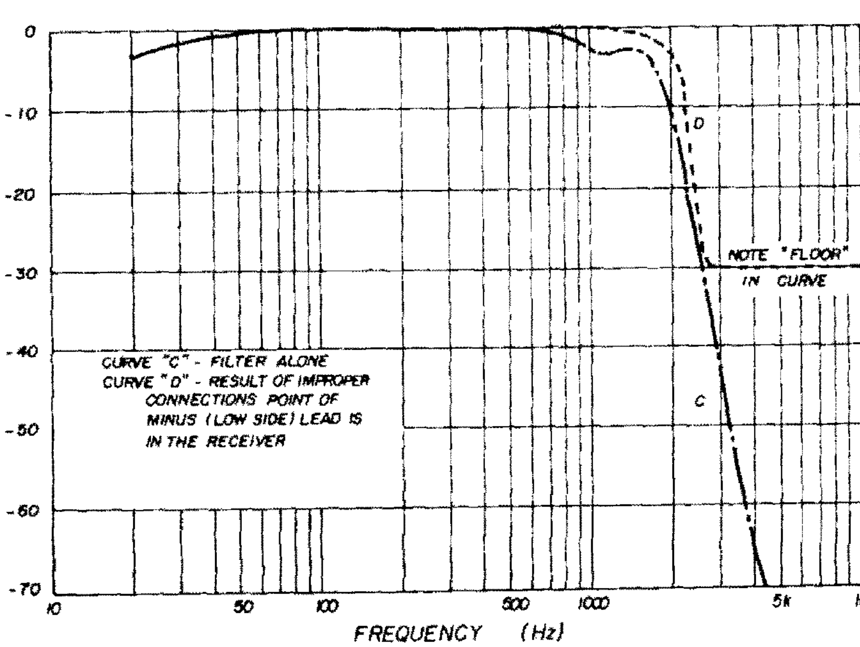


fig. 6. Example of a filter using off-the-shelf components. Also shown is the filter performance degradation due to improper grounding.

damage to the windings. A piece of insulating tubing or a wrap of plastic tape is sufficient. The 2-mF capacitors should connect to the switch with not much more than their lead lengths. The ground return from the switch should *not* go to the nearest chassis ground, but should return to the board, as *illustrated.* The interstage transformer and 500-ohm pot can be located at some distance from the board. In such a situation, it is best to use a shielded single-conductor cable with an outer jacket to prevent the shield from touching ground at any point except those selected. Miniature coax such as RG-174/U is excellent for this purpose.

ground loops

The foregoing precautions will help to prevent ground loops which can form paths around the filter. As a matter of fact, if your filter fails to do its job, it might be well to investigate this possibility before condemning the filter. **Fig. 6** is an excellent example of what can happen. I plotted curve C using a good lab setup. I installed the filter in an experimental receiver and it performed horribly. *Curve D showed that so much* signal was being fed around the filter that it was very ineffective. The culprit in this case was the low-side lead from the power supply. The point where it connected into the receiver was such that it formed a beautiful ground loop.

Fig. 7 illustrates a common trouble. The resistance marked R in that figure represents power supply impedance. For these receivers an adequate power supply can be the capacity multiplier job described by W6GXN in the February, 1970 *ham radio.*[3] A poorly regulated supply with an inadequate output filter capacitor may have an impedance measured in ohms. If connected to the receiver with a few long number 22 wires, the lead resistance *will* contribute *significantly to* power supply impedance. From **fig. 7** you can see that this power supply impedance is a common series element for both the detector and an audio amplifier. A small portion of the audio signal from the detector is fed around the filter by this

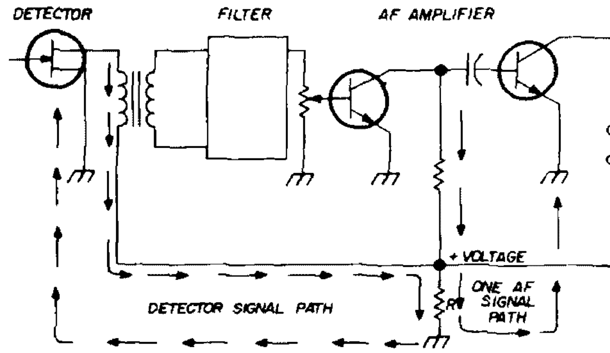


fig. 7. Example of audio filter performance degradation as a signal bypasses the filter by flowing through the common impedance of the power supply.

route, thus degrading filter performance. How much can be tolerated? If a weak signal reaching the first audio stage is 0.1 mV (not an impossible situation) filter degradation will begin to be felt with an unfiltered signal 80 dB below this

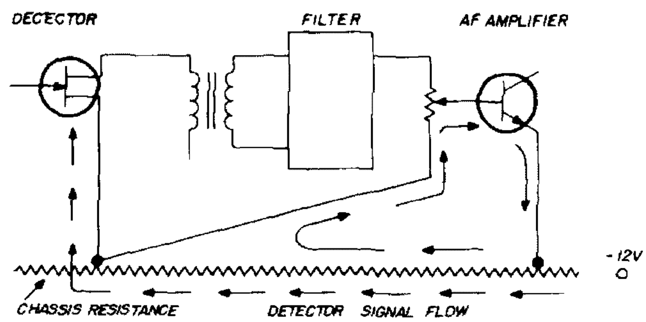


fig. 8. Filter performance is degraded by unexpected chassis resistances providing a signal path around a filter.

level – or 0.01 μV. In all probability, you won't notice any appreciable degradation at this level, but it is starting. A *one* microvolt sneak signal will produce a fair amount of degradation.

Fig. 8 shows another common mis-

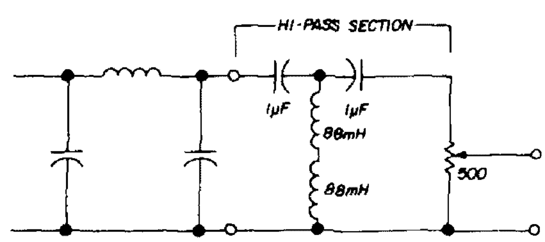
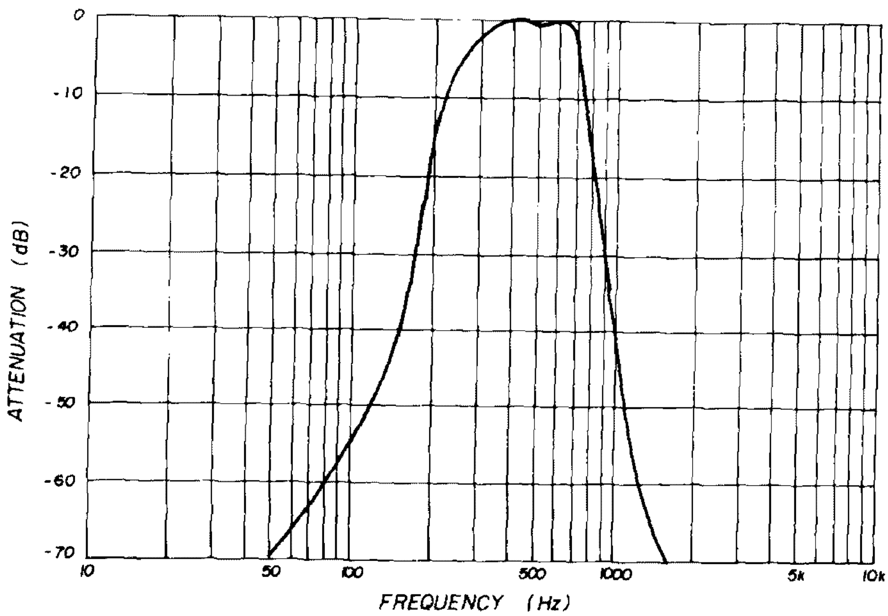
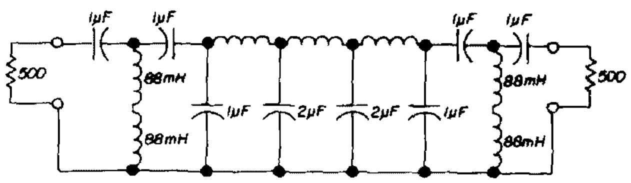


fig. 9. High-pass filter section that can be built from 88-mH toroids.



A



B

fig. 10. Extrapolated curve to approximate performance to be expected from a high-pass section at each end of a three-section low-pass filter. A typical filter of this type is shown in 10B.

take. The chassis here is represented as a resistance, which in actuality, it is. Particularly with aluminum, pressure connections (a solder lug bolted to a chassis) often show resistance due to the formation of oxides. Where dissimilar metals contact aluminum, oxidation is often accelerated, and aluminum oxide is an excellent insulator. **Fig. 8** shows that a part of the chassis resistance is common to both the detector and first audio stage. In this case, it could be that the volume control was located close to the detector to provide a symmetrical panel arrangement. Perhaps a person sees a ground point at the detector and thinks, "a ground is a ground," and fastens the volume-control ground wire there. As the illustration shows, the ground was merely a tap on a resistor.

Don't throw away that aluminum chassis, though! It will still do its job well if you follow a few simple rules. Try to make a straight-line layout of the stages. Avoid a U-shaped layout if possible, because that often brings high- and low-level ground points close together and leads to instability. In this case it can ruin filter performance. In a tiny 80- to 10-meter direct conversion receiver I built, it was necessary to locate the filter at one remote point and the volume control at another in order to cram the thing together. In a case like this, at audio frequency, it is possible to carry the ground through by using RG-174/U miniature coax. Both the hot and the ground leads were carried to the detector. No ground was allowed below decks on the filter board. The output went by coax to the volume control. Again, the ground side of the control was not physically grounded there. The coax led from the control back upstairs to the audio board. That is where the braid of the coax was allowed to go to ground. The result was a receiver with remarkable selectivity on CW. No filter degradation has been noticed. The last suggestion on grounding is to group the ground points associated with each stage at one point or close together. Don't include grounds for another stage at this

same point, if possible.

high-pass filter

The last item is fig. 9 which shows a high-pass filter section that can be built from the same 88-mH toroids. This section passes all frequencies *higher* than some selected cutoff frequency and can be used with the foregoing filters to suppress frequencies below about 250 Hz. Fig. 10 is not a measured curve, but is extrapolated from other measured curves to give a rough idea of what might be expected from a high-pass section at each end of a three section low-pass filter with the capacitor values indicated. A fairly substantial degree of mismtach has been deliberately introduced in the experimental filters here without much loss in usefulness.

If you want to try out a filter on an existing receiver having a headphone jack fed from a 4-ohm speaker line you could get a fallacious picture of expected performance. In such a case, since you would have audio power to spare, I would suggest connecting an appropriate terminating resistor *in series* with the hot side of the filter input to the hot side of the receiver phone jack. Similarly, if you are using high impedance phones, I would connect a resistor across the output of the filter to properly terminate it there, and then connect the headphones across that resistor so the parallel combination of the phones and resistor terminate the filter in a value roughly similar to that to be used a value roughly similar to that to be used in the final application. This sort of check will give a fair approximation of what to expect if it is built into something, and can serve as a method of determining if a bad feedthrough situation occurs in the finished project.

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2. William Wildenhein, W8YFB, "Regenerative Detectors and a Wideband Amplifier for Experimenters," *ham radio*, March, 1970, page 64.
3. Henry D. Olson, W6GXN, "A Survey of Solid-State Power Supplies," *ham radio*, February, 1970, page 38.

ham radio

n-way power dividers

and 3-dB hybrids

This article describes n-way power dividers and 3-dB hybrids, and shows you how to design them for hf and vhf

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Although amateurs who operate on the microwave frequencies are familiar with 3-dB hybrids and n-way power dividers, these devices are strangers to amateurs who operate on the lower bands. However, 3-dB hybrids have recently become available for use on vhf and uhf.

A 3-dB hybrid splits the input power equally between two outputs and provides more than 20-dB isolation between the output ports. An n-way divider splits the power n-ways (up to about 10 ways practically, I imagine) with similar isolation between output ports. The outputs of the n-way are in phase, while the outputs of the 3-dB hybrid are 90° out of phase.

Such devices are very useful in paralleling transistor rf amplifiers, for example, or in ssb systems where 90° phase shifts are necessary between two rf signals. By using lumped constant equivalents to the microwave transmission-line circuits, it is possible to design these devices for any amateur hf or vhf band. Circuit bandwidths are typically 10%. One circuit will, therefore, be useful over the full amateur band for which it was designed.

n-way divider

This is the most straightforward of the two devices (fig. 1). Power coming into the input port is divided equally between

the n output ports. Power coming in an output port is dissipated in the resistors connected to the common node rather than being reflected into the other outputs. This is the isolation mechanism that makes the circuit so attractive.

The design is accomplished with lumped-constant quarter-wave transmission lines (fig. 2). L and C values are determined by the characteristic impedance of the line (Z_o) which, in turn, is determined by the number of output ports. That is,

$$Z_o = \sqrt{NR_o}$$

where R_o = characteristic impedance of input and output lines and N = number of output ports.

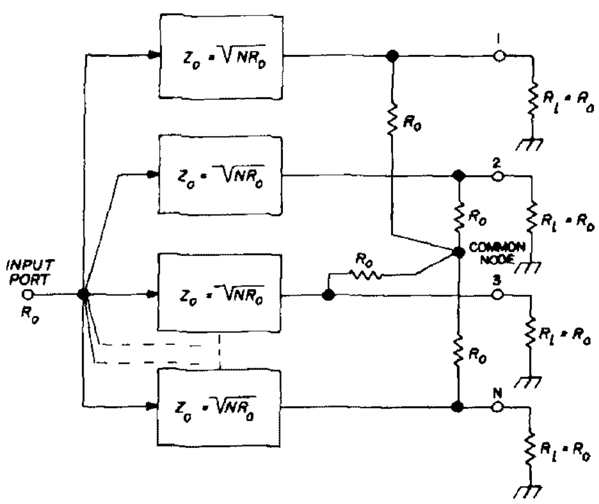

fig. 1. N-way power divider and combiner.

Fig. 3 shows a three-way divider for the 40-meter band (50-ohm transmission line). In this case,

$$Z_o = \sqrt{3 \times 50} = 86.5 \text{ ohms}$$

Therefore, in **fig. 2**, L = j86.5 and C = −j86.5 ohms. At 7.2 MHz, L = 1.6 μH and C = 250 pF. Use a reactance slide rule to determine these values (such as the ARRL R/L/C Calculator) or use the two reactance formulas:

$$L = 2\pi f X_L$$

$$C = \frac{1}{2\pi f X_C}$$

The closer you get to the calculated values, the more accurate the power division and the better the isolation. To obtain 30-dB isolation between output ports, you will have to be within 1% of the calculated values. Five percent com-

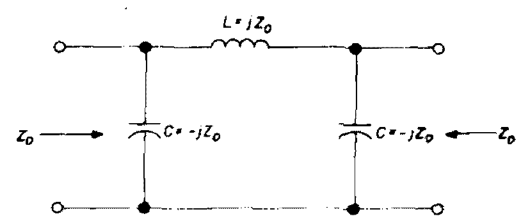

fig. 2. Lumped-constant quarterwave transmission line with a characteristic impedance, Zo.

ponent tolerances are probably the highest that can be tolerated without completely ruining output port isolation. Matching components between the paths is more important than obtaining precise values, so if you are unsure of your component tolerances, choose identical, if slightly inaccurate, values and sacrifice input impedance matching for improved isolation.

Turned around, an identical divider would be used to combine amplifier outputs. All the same isolation advantages apply. Total insertion loss through each divider (or combiner) is about 0.3 dB. Advantages of the n-way divider include equal power division from a straightforward circuit and good isolation between parallel amplifiers. Isolation at the signal frequency means fewer sneak paths to cause oscillation. Also, you can lose one paralleled amplifier without losing all output power.

hybrid

The 3-dB hybrid is more complex than the n-way power divider, but it is also more interesting and versatile. It may be used as a divide-by-two power divider in lieu of an n-way system and will provide similar isolation between the output ports. More interestingly, however, it can be used as a wideband phase shifter in systems requiring quadrature signals over a wide fre-

quency range. In a hybrid, the two output signals maintain a 90° phase relationship over a range of frequencies much wider than the useful 10% bandwidth which is determined from power matching considerations. Further properties will become apparent as we discuss the hybrid circuit.

Fig. 4 defines the basic hybrid properties. Power applied to input port 1 divides equally between ports 3 and 4. No power appears at port 2 (which must be terminated by a resistor equal to the characteristic line impedance). The signal at port 3 is delayed by 135°, the signal at port 4, by 45°, resulting in a 90° phase difference between the outputs.

This four-port network is completely symmetrical. Therefore, power applied to port 2 divides between ports 3 and 4 equally, with 135° delay to port 4 and 45° delay to port 3. Power applied to port 3 divides between ports 1 and 2 with no output at port 4, and so on.

If equal power is applied to ports 1 and 2, then equal power appears at ports 3 and 4 with 90° delay through the circuit. However, if the two equal input signals are 90° out of phase, then all the input power appears at either port 3 or 4 depending on whether the input to port 2 leads or lags the input to port 1.

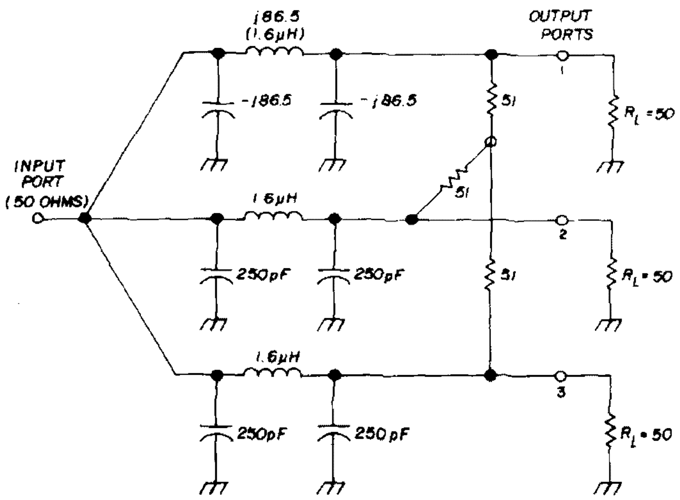

fig. 3. Three-way divider for operation at 72 MHz with 50-ohm transmission lines.

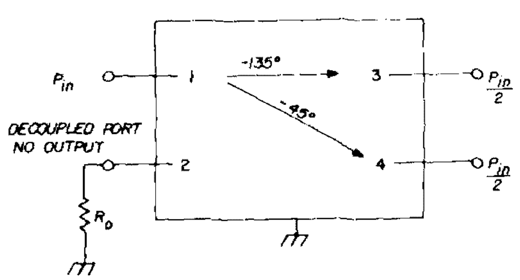

fig. 4. Basic circuit of the 3-dB hybrid.

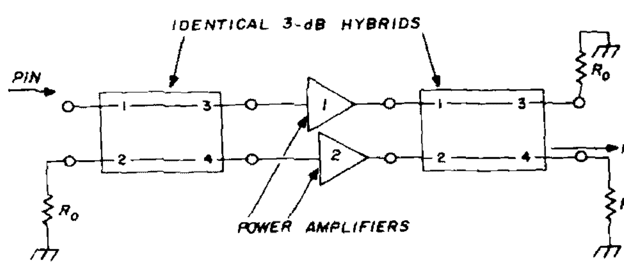

fig. 5. Using 3-dB hybrids as power dividers and combiners.

For power division and combining, a pair of hybrids is necessary (one at the input and one at the output) to maintain the proper phase relationships (fig. 5). Approximately the same signal frequency isolation between power amplifiers is obtained from the hybrid as from an n-way divider. All the power reflected back from amplifier 1 (due, say, to a mismatch) appears at ports 1 and 2 and

not at port 4 where it could feed amplifier 2. Similarly, on the output side none of amplifier 1's output appears at port 2 of the output hybrid, all of it drives the load. The insertion loss through each hybrid is 0.3 dB or about the same as the n-way divider.

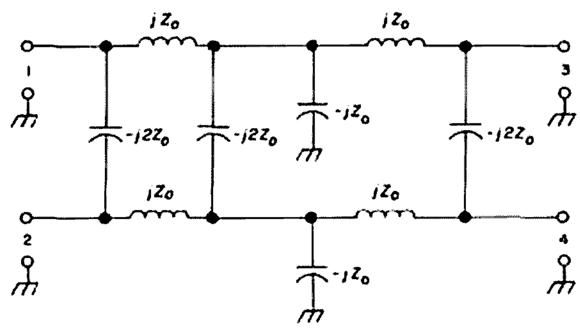

fig. 6. Basic circuit for a 3-dB hybrid for use with transmission lines with a characteristic impedance, Zo.

hybrid design

Hybrid design proceeds much the same as for the n-way divider. Fig. 6 shows the circuit and gives component values in terms of the characteristic impedance of the transmission lines with which it is used.

As a design example, a 3 dB hybrid for 50-ohm transmission lines will be developed for a center frequency of 14.25 MHz. I used this hybrid in a direct-conversion receiver so a good amount of test data is already available.[1]

From fig. 6 you can see that all reactances will be either Z_0 or $2Z_0$ (50 and 100 ohms in this case). These values are included in the circuit drawing in fig. 7. The component tolerances seem to be somewhat less critical here than in the n-way divider but you must still strive to obtain values as close as possible to those calculated. Component matching plays an important part in determining the ultimate isolation possible, so try to match components even if precise values are not available.

The hybrid for the ssb receiver was built with 10% inductors and 5% capacitors. The true center frequency was measured to be 14.45 MHz. At 14.25 MHz, the two output amplitudes were within 1 dB. Amplitude variation between the outputs around the true center frequency was 0.5 dB over ±0.3 MHz, which was acceptable. Over 10.3 MHz, the change in the 90° output phase relationship was not measurable. To provide for precise phase adjustment in the receiver, C_x (fig. 7) was replaced by 150 pF in parallel with the 100 pF trimmer. Amplitude compensation between the outputs was achieved elsewhere in the receiver. Power dividing and combining applications are not as critical as phase shifting and the trimmer might be omitted.

An interesting variation to the 3-dB coupler allows its use in swr measurements. By changing component values, it

"You can't get through a pileup like that . . . yuh gotta better chance out here on the edge."

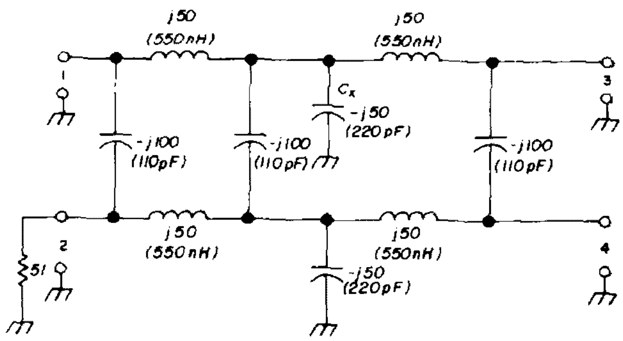

fig. 7. A 3-dB hybrid designed for 14.25 MHz operation with 50-ohm transmission lines.

is possible to change the coupling from 3 dB to 20 dB. With P_{in} applied to port 1 of this device, 99% is transferred through to port 3, and 1% to port 4. Since ports 3 and 4 are decoupled, the power at port 4 is only 1% of the forward power. Reflected power applied to port 3 *is treated* in the same fashion. Port 2 output is 1% of the reflected power only. These characteristics describe a directional coupler.

Fig. 8 shows the circuit values required for a 20-dB hybrid, and **fig. 9** shows the directional coupler application. Actual component values are determined from the circuit values as for the 3-dB hybrid.

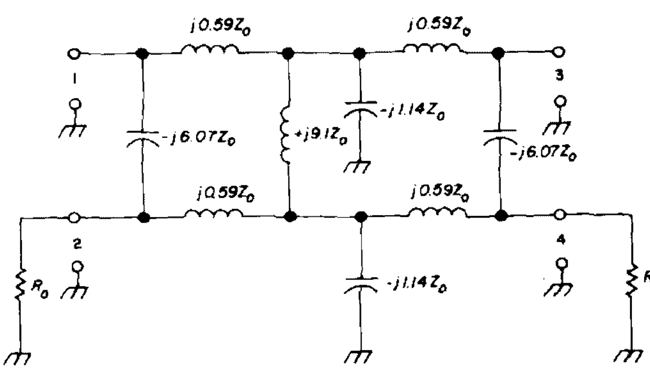

fig. 8. Reactance values for a 20-dB hybrid. These reactances can be converted into component values for any frequency of interest.

comments

Construct all of these devices symmetrically to minimize stray impedance differences between the paths. In fact, it is best to arrange them like the circuit diagrams. When paralleling power amplifiers, it is well to remember that these dividers provide isolation at the signal frequency only, and additional low-frequency isolation may be required in the power and signal leads to prevent destructive *low-frequency oscillation.*

Since all ports of these dividers are at transmission-line impedances, it is necessary to include matching networks in each amplifier. This is, in fact, an advantage since each individual amplifier can be tuned up in a 50- or 75-ohm system before you connect them in parallel. When they are paralleled, it will be necessary to retune each of them slightly

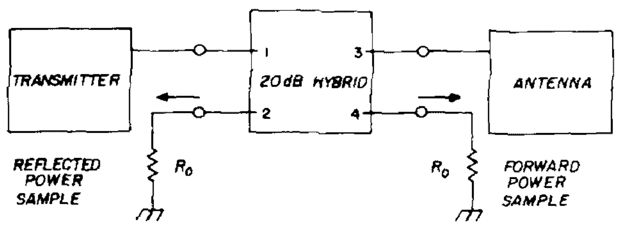

fig. 9. Using the 20-dB hybrid as a directional coupler for measuring swr.

to match phase shifts throughout the system. Choose one path as a reference and adjust the others to match it.

I have used n-way couplers in vhf solid-state transmitters to achieve output powers of 750 watts at 75 MHz and 650 watts at 225 MHz, the latter in a varactor tripler. Three-dB hybrids have been used in similar transmitter designs to achieve 1,000 watts at 100 MHz and also as *quadrature phasing* elements in ssb systems.

A computer program was the source of the hybrid component values. It was written by E. A. Johnston, K1YEY, and Bill Higgins of the staff of the MIT Center for Space Research in Cambridge, Massachusetts.

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ham radio

phase-shift RTTY monitor scope

By complete analysis of received signals, this unit can be the answer to RTTY tuning in heavy interference

Elmer E. Mooring, W3CIX, 9318 Millbrook Road, Ellicott City, Maryland 21043

A tuning indicator is almost a necessity to properly tune an RTTY signal. A cross-pattern scope[1] or the ST-5 and ST-6 type tuning meter[2,3] is fine for normal signals; however, with heavy interference, a phase-shift type scope[4,5] is perferable to help sort out the proper mark-space frequencies. The cross-pattern scope or tuning meter uses heavily-filtered descriminator signals from the demodulator for indication, thus filtering out much information about the received signals.

This phase shift monitor scope uses signals directly from the receiver output (or TU input) and displays a rotating line on the face of the CRT. The angle of rotation is a measure of frequency, while the length of trace indicates amplitude; thus, most of the CRT face area contains useful spectrum information. At a glance, you can determine approximate frequency shift of either the received or transmitted frequency, can tune the signal rapidly to correspond to the TU filters and can determine frequency separation of interfering signals. With a little practice, the scope pattern will aid in making the proper adjustment to produce usable copy from marginal signals.

theory

The heart of the phase-shift indicator is the simple RLC network shown in fig.

1. The series LC circuit *is* resonant near the mid frequency of interest. Near resonance, the impedance of the LC circuit is minimum; thus, E_T approaches zero while E_L is large and exhibits rapidly changing phase about the resonant frequency. Recalling oscilloscope display theory, two sine waves applied to the deflection plates of a CRT produce a pattern dependent upon the phase and amplitude relationship between these sine waves. With the two sine waves either in phase or 180° out of phase, a straight line appears whose angle on the CRT face depends upon the relative amplitudes of the two signals. With 90° phase shift between the two deflection plates, an ellipse appears with an eccentricity dependent upon the relative amplitudes between the two signals.

Referring to fig. 1, the LC circuit is resonant near the center of the frequency range of interest (2550 Hz). For a high-Q circuit, the phase angle varies from near 180° to 0° over a narrow frequency range, which yields straight lines on the oscilloscope tube. Around resonance, an ellipse appears since the signals are around 90° out of phase; however, the width of the ellipse is collapsed to zero since the amplitude of the series voltage (E_T) decreases to zero as the phase approaches 90°. Thus a straight line will appear on the CRT for all frequencies of interest and will rotate about the center of the tube face as the frequency varies.

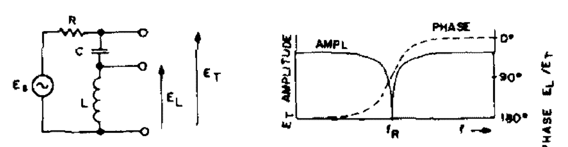


fig. 1. The basic RLC phase-shift network.

Typical patterns for a complete phase-shift monitor scope receiving RTTY signals appear in fig. 2. Most people prefer a *vertical line* at mid-frequency and clockwise rotation with increasing frequency. The CRT can be rotated in the mount to yield a horizontal or vertical pattern, and connections to one pair of deflection plates *can* be reversed to yield the desired direction of rotation.

circuit description

The complete schematic appears in fig. 3. The design is straightforward. The

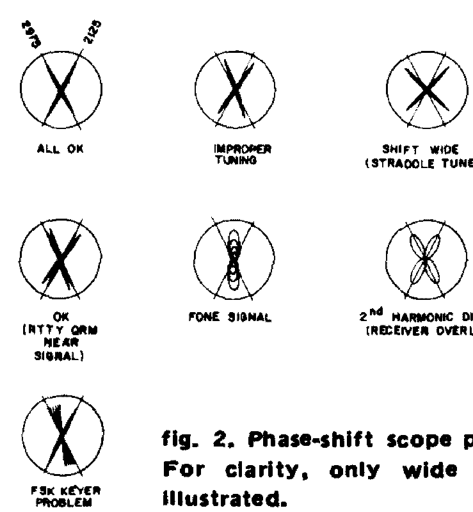


fig. 2. Phase-shift scope patterns. For clarity, only wide shift is illustrated.

input stage (Q1) yields a gain of about 20 (approximately the ratio R4/R5). R2 and R3 are bias resistors. Q2 is an isolation emitter follower which reduces Q1 collector loading and offers the current drive necessary for the LC phase-shift network.

The series compensating network (C2, L1 and R8) could be eliminated; however, it offers two advantages. C2-L1 resonate at approximately 2 kHz to yield an increasing impedance over the 2 to 3 kHz indication range which compensates for the increasing voltage across the phase-shift inductor (L2) with increasing frequency. This maintains the same scope trace length at 2125 Hz and 2975 Hz. Another advantage of C2-L1 is that signals far above and below the 2-3 kHz range do not produce a trace on the CRT, thus limiting the pattern to the desired range of frequencies.

R9, C3 and L2 comprise the phase-shift network described in fig. 1. C3/L2 resonate at 2550 Hz. Q3 serves as an isolating emitter follower to offer a high impedance across L2 and maintain high Q for the network. Q4 and Q5 are identical amplifier stages with a gain of approximately 20 (ratio of collector load to

emitter resistor). Resistors R14/R16 and R18/R19 furnish bias for the transistors.

construction

Parts layout is not critical. In the several prototype units built, I tried various construction techniques, all with equal success (point-to-point wiring, Vector board with push-in terminals, and tion of the steel. Of the several prototype units I built, no trace hum was experienced with the exception of an early prototype in which the power transformer touched the CRT socket. This hum was eliminated with a cut-down CRT neck shield (Heathkit 206-180).

You can use a standard CRT bezel to dress up the front panel and for mounting

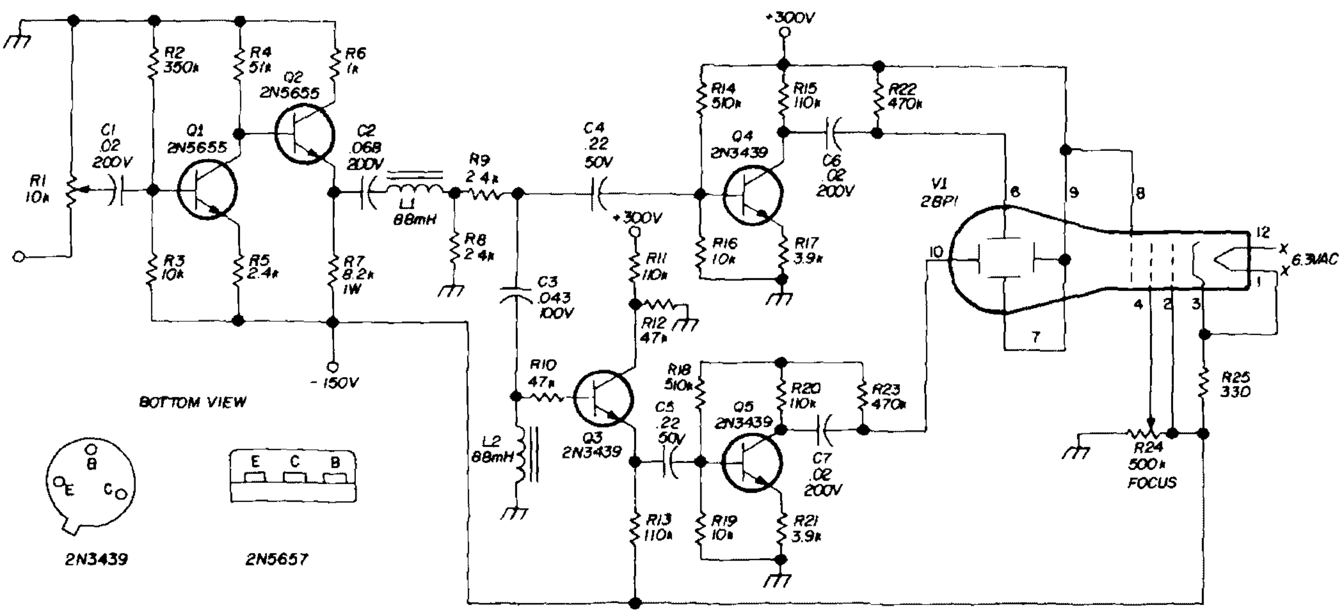


fig. 3. Schematic of the transistorized RTTY monitor scope. Q1 and Q2 may be the same as Q3, Q4 and Q5; there are many substitutes for these transistors. In choosing V1, try to get a CRT with P1 phosphor.

Vector board with Circuit-Stik). Preferably, the power transformer should be mounted to the rear of (and away from) the CRT, and the input amplifier (Q1) should be towards the front of the unit. The only shielded wire is from the input jack to the gain control (R1).

For the two inch CRTs, a 3 x 5 x 9½-inch aluminum chassis is adequate to house the CRT and circuit board. Use an aluminum cover for the hooded top, or alternately, build the unit with the chassis open side down and use a flat aluminum bottom plate. If the power supply is to be built-in, select a chassis size to prevent crowding the power-supply components around the CRT. Otherwise CRT shielding may be required to prevent trace hum. Do not use a steel chassis or bottom plate, since the centering would tend to shift with magnetiza-

the CRT. The unit shown uses an old meter case with the movement and glass removed and the opening filed round. A rubber O-ring slipped into the modified meter case provides a shock mount for the CRT face. The tube neck is supported with an aluminum clamp and spacers. Save space and money by omitting a special tube socket for the CRT. Push pins from a standard old tube socket, solder them directly to the leads and push spaghetti over the pins for insulation (octal socket pins fit the 2BP1, larger pins are required for the 3AP1).

Circuit components are not critical, so feel free to use the junk box; however, C2 and C3 should be high quality (e. g. Sprague Orange Drop) to yield high Q and to minimize calibration drift. Of course, it is an unusual junk box that contains 400 V breakdown transistors. C6

and C7 should have low dc leakage so as to not affect CRT centering. Although not necessary, the gain and Q can be optimized by selecting high gain (h_{FE}) transistors for Q2 and Q3. The circuit in fig. 4 can be used to select the two highest gain transistors from a batch for those persons using the same type throughout. The lower the measured

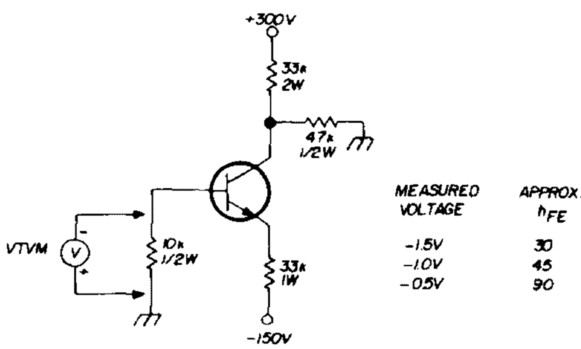


fig. 4. Test circuit for selecting high-gain transistors for Q2 and Q3. The lower the negative base voltage, the higher the Q.

(negative) voltage at the transistor base, the higher the gain. You can use a 20-kilohm-per-volt (or greater) vom in place of the vtvm for selection; however, the h_{FE} values listed in fig. 4 will be inaccurate (lower voltage reading for equivalent h_{FE}). For Q2 and Q3, just choose the units with the lowest measured base voltage.

Minimal power is required and voltages are not critical, thus power for the unit can be taken from other station equipment. The 300-volt load 6 mA and the -150 volt load draws 12 to 14 mA (including CRT). For a built in supply, I used the smallest transformer on hand, a 250 V ct unit and the circuit shown in fig. 5.

Almost any small CRT can be used, such as the 2BP1, 2AP1, 3BP1, 3AP1 or 3RP1. This power supply and circuit has adequate output swing to drive any of them, since the CRT deflection factor is greater with reduced accelerating potential (usually 1000 to 1500 volts). All are sufficiently bright for a well lighted room.

Be certain to check pin connections and heater requirements for your tube (3AP1 requires 2.5 volts). Don't worry about the spot that finally burns on the center of the tube face during undeflected periods; no useful information is contained in the center of the tube anyway. Centering controls are dispensed with for simplicity. Most CRTs are fairly well centered as is; however, a small magnet can be glued to the CRT neck for centering if desired. A small chip from a dime store magnet is adequate, just slide it around the tube neck until the display is centered on the tube face, and a spot of epoxy or Q-dope will hold it. Flexible black-board magnets are ideal for this purpose.

The phase-shift circuit can be used with a standard test oscilloscope for RTTY monitoring; however, connect the output directly to the deflection plates and not through the scope amplifiers. The additional amplifier phase shift would distort the straight-line pattern.

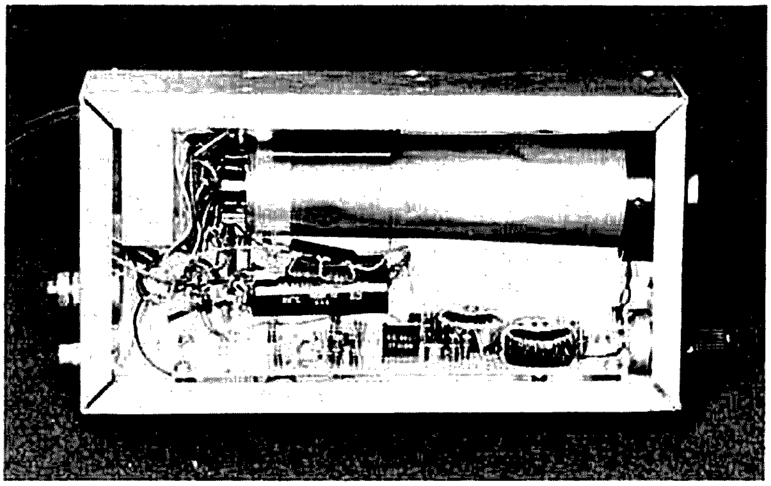
initial adjustment

At first turn-on, defocus the spot until after the centering magnet is glued on. With power applied and the input sensitivity control (R1) turned down, the following dc voltages should exist:

Q1 collector	≈ approximately one half of the negative supply voltage
Q2 emitter	≈ within 1 volt of Q1 collector
Q3 emitter	= -0.5 to -1.0 volts
Q4 collector Q5 collector	= approximately one-half the positive supply voltage

If the readings are off more than 20%, adjust the operating points by changing the value of R2, R14 and R18. Increasing R2 reduces the negative voltage at Q1 collector and Q2 emitter. Increasing R14 increases Q4 collector voltage, increasing R18 increases Q5 collector voltage.

Apply a 2 to 3 kHz sine wave at the input. A straight line should appear on the scope; adjust the amplitude for three-quarter screen deflection. Focus the spot



Interior view of an early prototype using point to point wiring on Vectorboard. Note the power transformer crammed against the CRT socket and filters near the neck of the CRT — both necessitated the neck shielding shown to eliminate a slight trace hum. Input connector and focus pot are visible at the rear, sensitivity pot and power switch are on the front panel.

with R24. As the frequency is varied, the line should rotate about the center. If you prefer rotation in the other direction, reverse the deflection plate connections (pins 6 and 7) at the CRT (or the equivalent pins for your CRT). The line should rotate smoothly from 2 to 3 kHz, with reduced angular rotation and shrinking in size below 1 kHz and above 4 kHz.

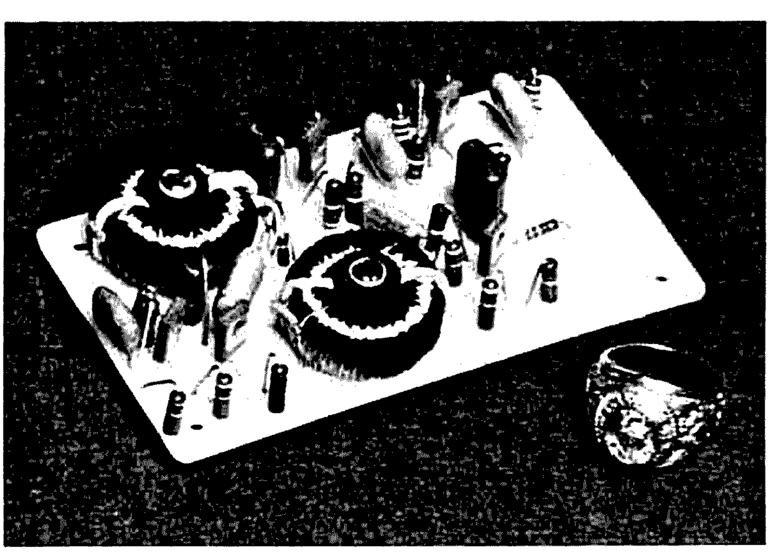
Calibration marks are placed on the face of the tube by stretching fine black threads across the tube face (behind the bezel) and adjusting these over the trace at desired frequencies (2125, 2295, 2975 Hz). A spot of Q-dope out of sight at the tube edge will hold the threads in place. If available, a calibrated oscillator or an oscillator and counter may be used for calibration.

For those using the ST-5 or ST-6 demodulator, the calibration threads can be easily placed right on the filter frequencies. Just connect the monitor scope to the receiver output, turn on the calibrator and bfo and tune for a tone. Tune the bfo for maximum on the TU tuning meter (or maximum measured discriminator test point voltage), and repeat for mark and space on the narrow and wide shift positions of the TU. Adjust the calibration threads over the scope pattern when you obtain the peak discriminator output.

parts

The junkbox should supply most resistors and capacitors, but few will have CRTs and 400-volt breakdown transistors.

The power transformer can be a Stancor PA-8416, $4.66 from Lafayette or the Thordarson 22R39, $4.37 from Newark. These transformers have a 6.3-volt heater supply which is suitable



The complete phase-shift circuit without power supply. This unit was built using Circuit-Stik and the transistors used were 2N5657 rather than 2N3439 shown on the schematic.

for the 2BP1 and other CRTs. For the 3AP1, you will need a 2.5-volt heater supply. Van's (W2DLT) 302 Passaic Avenue, Stirling, New Jersey 07980, offers unused 3AP1 CRTs in original boxes for \$4.50, first-class postpaid anywhere in the United States. Van also has 88-mH toroids at five for \$2.00. 2N3439 transistors are available for \$1.19 each from General Radio Supply Company, 600 Penn Street, Camden, New Jersey 08102. M. Weinschenker, K3DPJ, Box 353 Irwin, Pennsylvania 15642 offers .22 μF 100-volt mylar capacitors at 16 for \$1, and .02 μF 200-volt dipped capacitors at 20 for \$1, postpaid.

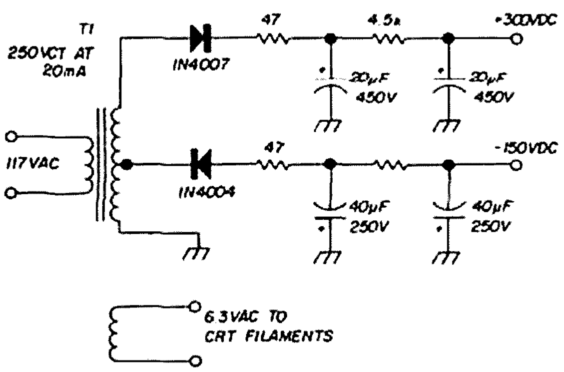


fig. 5. Suitable power supply for the monitor scope.

I wish to thank Bruce Meyer, WØHZR, for his original vacuum-tube phase-shift scope design which served as the foundation for this design. I also want to thank Don, WA6PIR, for urging me to write this article to share the design with other RTTY enthusiasts.

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crystal oscillator frequency adjustment

The frequency of crystal oscillators is often adjusted with a trimmer capacitor — this article explains how much adjustment can be expected

Calvin Sondgeroth, W9ZTK, 800 Fifth Avenue, Mendota, Illinois 61342

Because of its inherent stability in oscillator circuits as a frequency-control component, the quartz crystal finds much use in circuits where frequency tolerance and stability are important to proper circuit operation. Present transmitter and receiver design trends make good use of this fine frequency-control element.

Much has been said about the necessity for operating the crystal in a circuit suitable to produce the required nominal operating frequency. Although the crystal has an extremely high Q factor and good stability with temperature changes, the circuit in which it is used can, and does, have some measure of control of the final frequency of the overall oscillator circuit. This is because of the ultimate load capacitance across the crystal terminals in the circuit it is plugged into. For this reason, crystal manufacturers specify this operating load capacitance to ensure that the crystal will control an oscillator within their manufacturing tolerances and specifications.

Of course, good use can be made of the small effect load capacitance has on the operating frequency. You can make slight frequency adjustments by changing the load capacitance with a small trimmer capacitor somewhere in the oscillator circuit. Receivers and transmitters can be set on frequency in this manner. Another use of this frequency change is in RTTY applications where the oscillator frequency is changed for frequency-shift keying.

While these frequency adjustments are often used, not much is usually known in advance as to what magnitude of frequency change can be expected for a given change in load capacitance. All too little has been published in this area, and manufacturers often do not specify this sort of data for their crystals. There are, however, some rather simple relations that apply to most crystals common to amateur use that can be used to get at least a better idea of the frequency changes involved. This is good information to have available, as it allows the selection of a crystal which will produce the desired nominal frequency as well as the necessary adjustment range for final frequency calibration. Also, it allows a lesser tolerance crystal to be selected; variations are compensated for by trim-

ming the frequency. A lower tolerance crystal is usually much less costly than one with tighter specifications.

crystal equivalent circuit

In the course of developing crystals and associated oscillator circuits, early researchers came up with an equivalent *circuit for the* crystal. This familiar circuit and its impedance vs frequency curve is shown in **fig. 1.** The circuit consists of an inductance, capacitance and resistance in series with the entire combination shunted by another capacitance, C_o. L_m in **fig. 1** is the equivalent inductance of *the crystal,* C_m *is the equivalent* capacitance, and R_m is the effective series resistance. C_o is a static capacitor formed by the crystal electrodes and the stray capacitance of the holder leads and terminals.

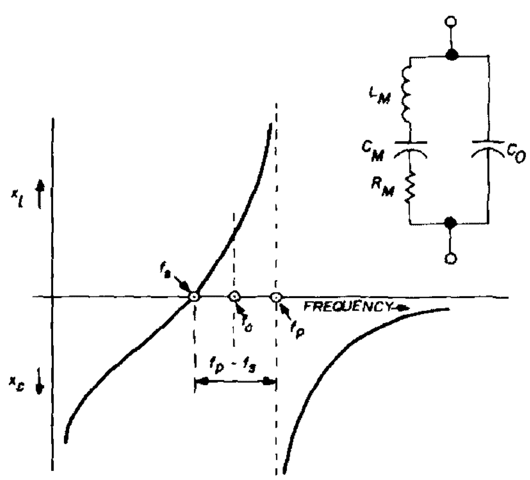

fig. 1. Crystal equivalent circuit and impedance vs frequency curve.

The thing that makes the crystal so stable in an oscillator circuit is that the motional inductance, L_m, is measured in terms of hundreds of millihenries and has an equivalent Q of 20,000 to 200,000. The coil in an ordinary tuned circuit might have an inductance measured in microhenries with a Q of only 100 or less at the same frequency.

The motional capacitance, C_m, has an extremely small equivalent capacitance in order to be resonant with L_m at the crystal frequency. These parameters are termed *motional* inductance and capacitance because they are equivalent values associated with the actual molecular vibrations in the quartz.

Inspection of the circuit in **fig. 1** indicates that the crystal has two points of resonance, one of which is the frequency of L_m and C_m in series. (This is an approximation, close *enough for* amateur purposes.) At this series-resonant frequency, indicated as f_s on the impedance curve, the crystal looks like a small equivalent resistance, R_m, which is determined by the mounting and preparation of the crystal in its holder.

As the frequency is increased above f_s, the overall series arm of the circuit takes on the character of an inductance; at some slightly higher frequency the overall circuit, consisting of the equivalent inductive reactance of the series arm and the electrode/holder capacitance, C_o, in parallel, becomes resonant. At this point, the crystal looks like a parallel tuned circuit with a very high Q and resistive impedance. This frequency is called the parallel or anti-resonant frequency and is indicated by f_p in **fig. 1.**

A few oscillator circuits use the series-resonant frequency for operation, but they are usually confined to applications at extremely high or low frequencies. Most of the oscillator circuits found in amateur equipment, Pierce, Colpitts, etc., use the crystal in a parallel-resonant fashion. Since there is some load capacitance offered by the oscillator circuit, the actual operating frequency, f_o, comes out at a point that is between the two natural crystal *frequencies,* f_s *and* f_p.

The following discussion deals with relations that can be used to determine what this operating point will be in terms of approximate crystal parameters and added oscillator load capacitance.

capacitance ratio

An important crystal parameter that is not often mentioned or specified by the manufacturer is one called the *ratio of capacitances.* This is simply the ratio of the crystal holder static capacitance, C_o,

to the crystal motional capacitance, C_m. Since the motional capacitance is very, very small in terms of "real" capacitors, its value for most high-frequency AT-cut crystals does not change much over a fairly wide frequency range. At least its change is very small with respect to the rather large (in comparison) value of C_o.

The holder capacitance is usually 5 pF or so, while the motional capacitance is often measured in thousandths of a picofarad. Thus, the ratio of capacitances is relatively constant over a wide range of frequencies, and is primarily a function of the type of crystal cut. It would be well to repeat that this discussion applies only to fundamental-mode AT-cut crystals. This is the type most common in high-frequency amateur equipment.

The capacitance ratio, as it is often called, turns out to have a value between two and three hundred. For experimental purposes, you might consider it to be 250. There is a very simple relationship between this capacitance ratio and the natural bandwidth of the crystal. This bandwidth is defined as the frequency spread between the series and parallel resonant frequencies. It is equal to f_p-f_s, and can be expressed as:

$$f_p - f_s = \frac{f_p}{2rc} \qquad (1)$$

If you take the capacitance ratio, r_c, to be 250, **equation 1** shows that to find the natural bandwidth for a crystal you divide the crystal frequency by $2r_c$ or 500. It should be noted that this relationship indicates that, for any crystal, the relative bandwidth in terms of percentage of the nominal frequency is the same. For a capacitance ratio of 250, you can expect a bandwidth of the order of 2000 parts per million for any frequency. As you already know, the higher you go in frequency, the more you can trim your oscillators. **Equation 1** gives an indication of why this is true.

As an example, you might have an eighty-meter, fundamental AT-cut crystal at 3.600 MHz. Using **equation 1** you find that its natural bandwidth comes out to be 3,600,000 divided by 500 or about 7200 Hz. This gives a rough idea of the frequency adjustment possible with this crystal, although, as you shall see, the actual amount of adjustment becomes quite a bit less as oscillator load capacitance is added to the crystal.

effects of load capacitance

In an oscillator circuit of practical design, the crystal becomes loaded with additional capacitance which appears in shunt with the holder capacitance, C_o. If this additional capacitance is designated C_x, the following expression gives the amount of frequency spread between the crystal's series resonant frequency, f_s, and the new parallel resonant frequency of the oscillator, f_o.

$$f_o - f_s = \frac{f_o}{2r_c} \cdot \frac{1}{1 + \frac{cx}{co}} \qquad (2)$$

From **equation 2** you can see that the amount the oscillator frequency is above the series resonant frequency is some fraction of the total crystal bandwidth given by **equation 1.** The fractional multiplier is a function of the ratio of the added load capacitance to the holder capacitance, C_o. This relationship indicates why the actual amount of frequency adjustment when load capacitance has been added is considerably less than the overall bandwidth of the crystal. To build a practical oscillator circuit, it is necessary to add considerable load capacitance, C_x, and the manufacturer's specifications often require on the order of 32 pF for nominal frequency tolerance to apply.

Since the holder capacitance is about 4 or 5 pF, the difference between the series resonant frequency and the operating frequency of the oscillator is somewhere around 1/6 to 1/8 the total crystal bandwidth. For the 3.6-MHz example above, the oscillator might be operating as little as 1000 Hz above the series-resonant frequency. Further investigation of **equation 2** indicates that even though you might add considerable capacitance above 32 pF, the actual frequency change will be quite small.

To use **equation 2** it is convenient to set up the load capacitance values as integral multiples of C_o. For HC6/U crystals, the holder capacitance is usually quite close to the 5 pF figure mentioned. This is due to the fact that the electrodes are electro-deposited on the quartz blank and are very close to the same size for any frequency. Likewise, thickness dimensions do not vary too greatly.

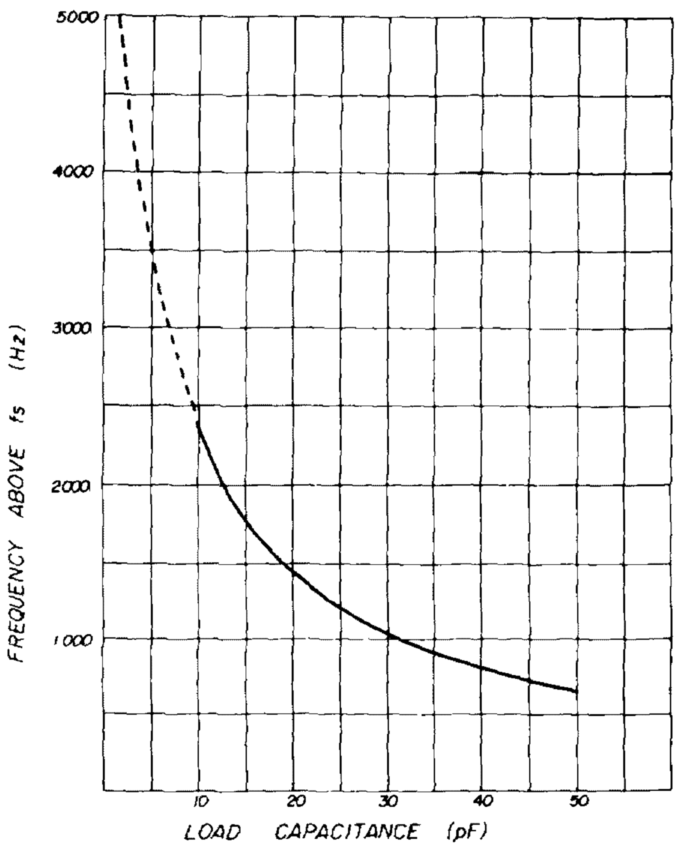

fig. 2. Effect of load capacitance upon the operating frequency of a 3.6-MHz AT-cut crystal. *For this graph,* **C_o = 5 pF, capacitance ratio = 250, f_o = 3.6 MHz.**

For FT-243 pressure-mounted crystals, the static holder capacitance can be estimated by calculating it from the blank dimensions since the crystal is easily disassembled. The dielectric constant for quartz is about 4.3 and these crystals usually have somewhat higher holder capacitance – on the order of 8 or 9 pF. Thus, it can be seen that they are more difficult to shift in frequency than their wire-mounted plated-blank counterparts. (Their shunt capacitance, and thus, capacitance ratio, is higher.)

practical values

A graph plotted for **equation 2** is shown in **fig. 2.** The frequency changes on the vertical axis are for the 3.6-MHz crystal used as an example above, and values of load capacitance are plotted along the horizontal axis. The portion of the curve below 10 pF is dotted. This area is *normally unusable in practical* oscillator circuits since a minimum load must be presented to the crystal to establish the circuit parameters suitable for oscillation. This is why the oscillator frequently quits when you try to turn the trimmer too far out in an attempt to get it on frequency.

There is, of course, some limit to the maximum capacitance that is suitable, but by the time this value is reached, the effects of capacitance change on frequency are practically nil. In fact, by the time you get the circuit oscillating with 10 pF across the crystal, you are down in frequency from f_p by roughly 2/3 the crystal's natural bandwidth. Further increases in load produce less and less change in frequency.

This is one of the reasons why 32 pF has become somewhat of a standard for crystal calibration. This amount of load minimizes changes in frequency that might occur with unwanted capacitance changes vs temperature, shock and the like. Nevertheless, a 32-pF load allows a nominal range of adjustment for final frequency calibration.

It should be pointed out that **fig. 2** is plotted for a holder capacitance of 5 pF and a capacitance ratio of 250. Actual values will be somewhat different for different crystals, but these are good figures for estimating how much frequency adjustment can be expected at a given nominal operating frequency.

Going back to the 3.6-MHz crystal as an example, you can see from **fig. 2** that if it is going to be on nominal frequency at say, 30 pF load capacitance, you might expect an adjustment range of about 1680 Hz. However, in going from 30 pF to 50 pF, you would get a change of about 360 Hz, while going the other way from 30 pF down to 10 pF, would

produce a change of some 1320 Hz. This indicates quite clearly what amateurs *have frequently taken advantage of — it's* easier to move a crystal up in frequency with the trimmer than down.

For some circuits, it might be well to obtain crystals which are a bit low (specified at 32 pF) and depend on the trimmer to move them up on frequency. The equations and **fig. 2** can be used to get a fair idea of where the nominal frequency should be specified.

Bear in mind however, that the data presented here is based on a couple of approximations; i. e., assumed values for the capacitance ratio and holder capacitance. The latter can usually be calculated or measured to within rather close limits. And for "worst case" calculations, a value of 300 for the capacitance ratio might be a better value to use. This would ensure that adequate frequency adjustment range would be available for a given crystal.

actual measurements

As a rough check on some of the relations presented here, I set up the oscillator circuit shown in **fig. 3**. The equivalent load capacitance offered to the crystal is about 20 to 35 pF as the 30-pF trimmer is adjusted through its range. To simulate the 3.6 MHz example calculations above, three HC6/U crystals in the 80-meter band (3.5 to 3.6 MHz) were tested. For a capacitance change of 20 to 35 pF, the measured frequency changes were 360, 455 and 510 Hz, indicating a fairly wide spread among different crystals at the same *approximate* frequency. **Equation 2** shows that slightly more (about 550 Hz) adjustment should be possible, indicating that a value of 300 for the capacitance ratio would probably be a better choice. In addition, it would be well to try to obtain a value for the holder capacitance — either from the manufacturer or by actual measurement. These crystals *measured* very close to 5 pF C_0.

conclusions

Because of its large value of motional inductance and extremely high Q factor, the crystal is a good component to use as a *frequency-control element in oscillators.* It is interesting to note that the impedance vs frequency curve in **fig. 1** is drawn out of scale. Below the series

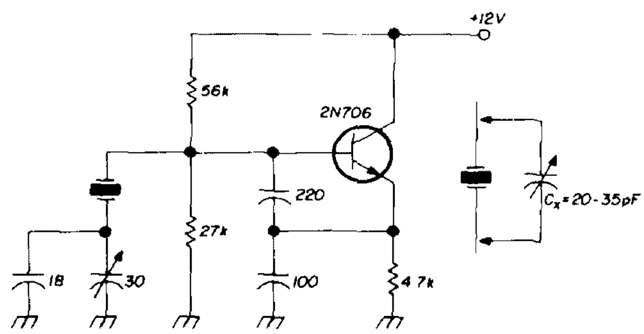

fig. 3. Oscillator test circuit.

resonant frequency of the crystal the capacitive reactance (of the holder capacitance) is much smaller than the reactance values between series resonance and the parallel-resonant frequency.

Likewise, the frequency scale is expanded a great deal around the crystal frequency to indicate the crystal's bandwidth. If the reactance axis were drawn to scale, the inductive reactance approaching parallel resonance would go off the paper. It is this rapid change in reactance with minute frequency changes that gives the crystal its ability as an excellent frequency controller.

Although the crystal is extremely stable, small changes in load capacitance in the oscillator circuits can provide for slight *adjustment* of the oscillator frequency. **Equations 1** and **2** show the amount of adjustment that can be obtained once some crystal parameters have been either measured or estimated.

While these calculations are not exact, they give good approximations for the amounts involved for a given frequency of operation. The accuracy, of course, is limited by the inability to establish true values for these parameters. With exact values, the relations in **equations 1** and **2** are exact.

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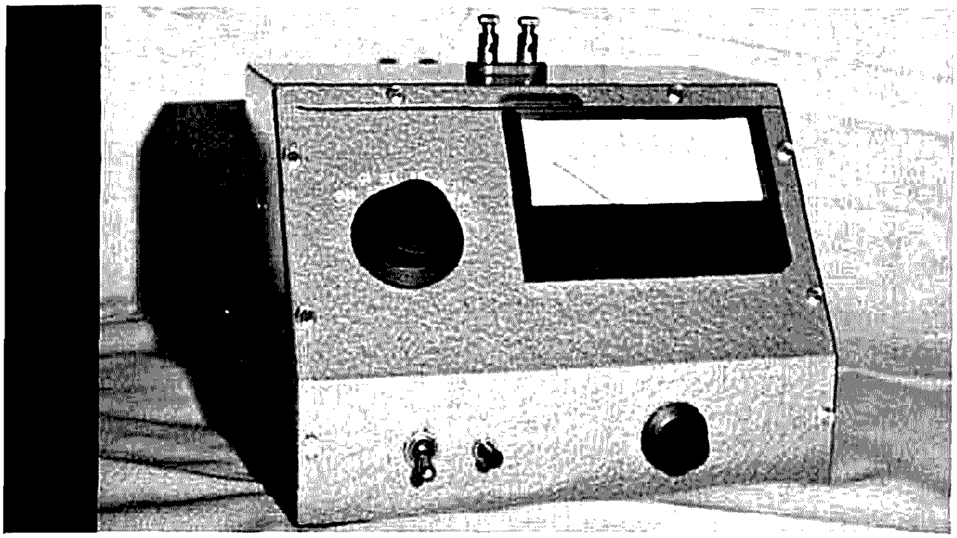


direct-reading capacitance meter

Conservative design and compact packaging are featured in this useful addition to your test bench

R. W. Johnson, W6MUR, 7511 (9) Clay Avenue, Huntington Beach, California

A direct-reading capacitance meter is a particularly useful test instrument. With it you can check capacitors whose markings have worn off or whose tolerance is unknown. Many ceramic capacitors, for example, are GMV (guaranteed minimum value), and their actual value may be as much as ten times the value marked. A direct-reading capacitance meter is also more convenient to use than a bridge when determining capacitance changes with temperature.

The instrument described can measure capacitance in five decades, from 100 pF through 10 μF, to an accuracy determined mainly by the calibration standard used and the accuracy of the meter movement. Five percent accuracy is easily attainable. The parts cost is under $20. The instrument will operate from any unregulated power supply reasonably free of ripple and capable of delivering 28 volts at 50 mA.

block diagram

A reference sawtooth oscillator (**fig. 1**) determines the over-all period of a pulse whose width is linearly proportional to the unknown capacitance, C_X. A sawtooth oscillator was chosen because of

the simplicity of switching ranges using only one RC time constant. Through an emitter follower (to avoid loading the reference capacitor, C_R), the sawtooth operates a Schmitt trigger, which changes state at a fixed and repeatable voltage level part way up on the sawtooth as the reference capacitor charges. A shaper-driver operates a discharge transistor to discharge the unknown capacitor, C_X, or the other of the two inputs is positive and no gate output occurs. As soon as the second Schmitt trigger, ST_2, fires part way up on the charging waveform for C_X, the NAND gate output again returns to zero.

Thus the time between leading edges of the NAND gate output is the same as the period of the reference oscillator, and the width of the NAND gate output is

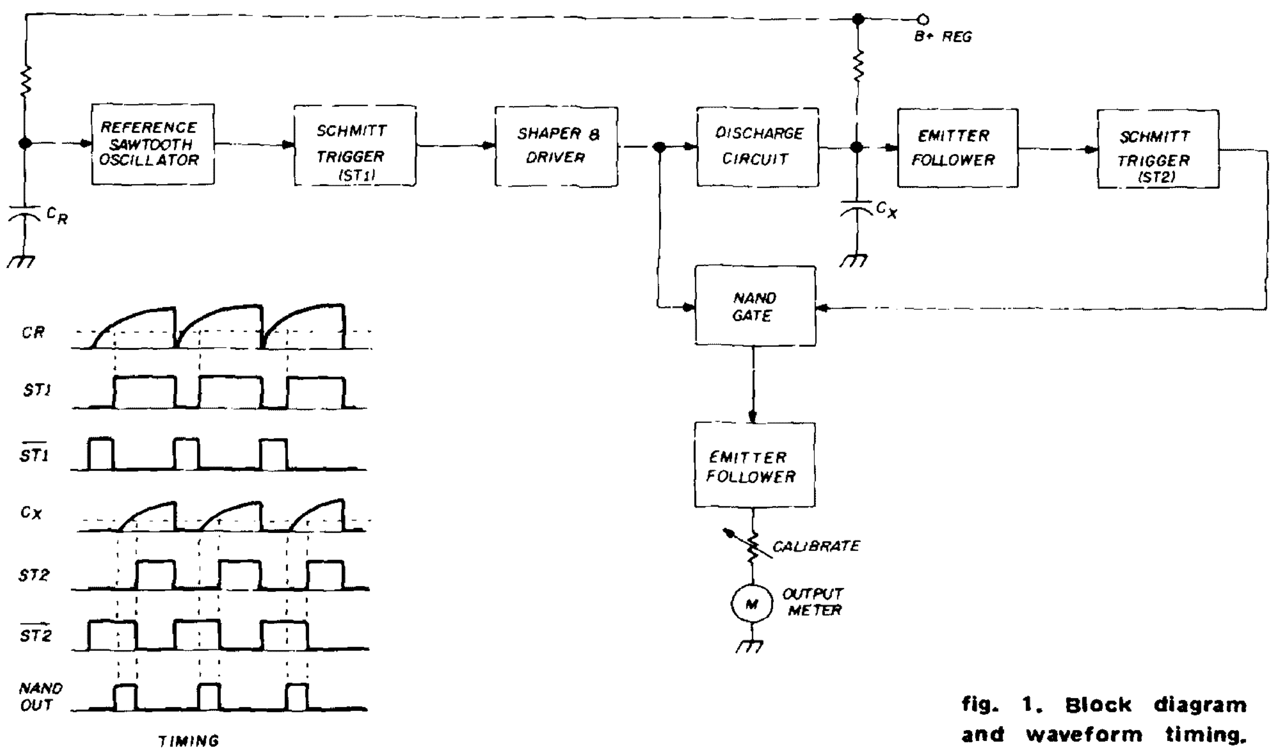


fig. 1. Block diagram and waveform timing.

and also furnishes one input of a NAND gate. The charging voltage across C_X is applied through an emitter follower to another Schmitt trigger, whose output is the second input of the NAND gate.

C_R is the reference capacitor of the sawtooth oscillator. When the first Schmitt trigger, ST_1, resets at the end of the reference period, the unknown capacitor is discharged and clamped by the discharge circuit at almost zero volts until ST_1 triggers again part way up on the reference sawtooth (see timing diagram). At this time the unknown capacitor, C_X, charges through a precision resistor of value switched by a range switch. During this C_X charging *period (only) an output* from the NAND gate occurs since both of its inputs are negative. At other times one

linearly proportional to capacitance C_X. If this pulse is applied to an average-reading meter, the meter will read a current linearly proportional to the unknown capacitor, C_X. No special scale is required on the meter.

design considerations

The main reason for using the NAND gate over other possibilities was to make the meter read zero when C_X is disconnected; the ST_2 circuit is triggered for this condition.

Fig. 2 is the schematic for the capacitance meter. While this circuit seems to have a lot of transistors, it should be remembered that the RCA CA3046 IC array is used, which has five transistors in a single 14-pin dual in-line package. So

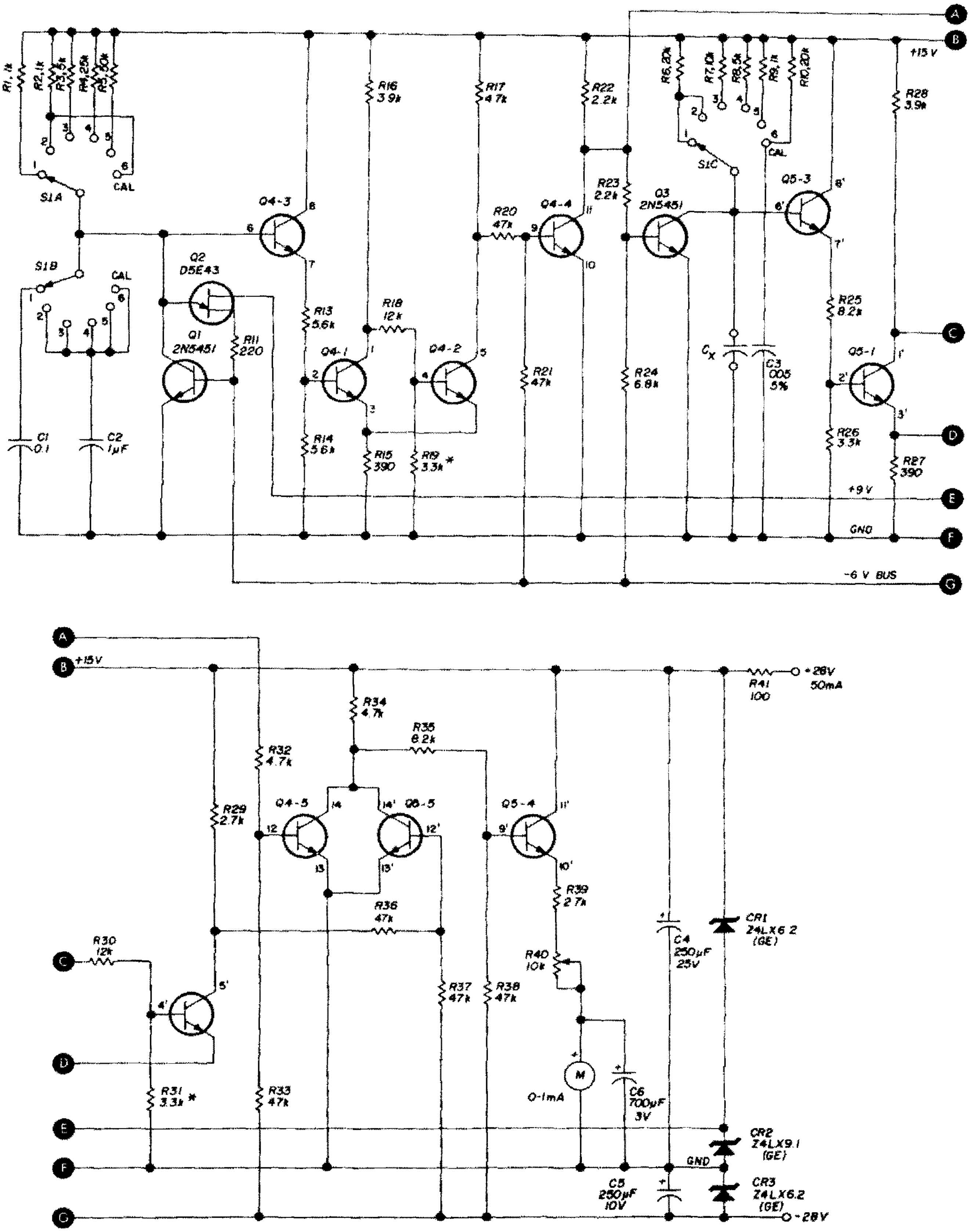


fig. 2. Circuit schematic. RCA type CA3046 IC arrays are used for Q4, Q5, resulting in compact packaging. Resistors R1 through R10 are 1% tolerance. R26 is 3.3k nominal (may be 2.7k or 3.9k). Sometimes it is necessary to change R19 to 12k to make the first Schmitt trigger work. Also, due to variations in the components used, changing R7 to 9720 ohms and R8 to 4875 ohms will probably improve accuracy on ranges 3 and 4. R19 and R31 may need to be increased because of CA3046 gain.

ten of the transistors are simply two CA3046s, labeled Q4 and Q5 in **fig. 2.**

One of the more troublesome problems in developing this capacitance meter was getting a good unijunction sawtooth oscillator without having to resort to expensive UJTs such as the 2N494. In the ordinary UJT sawtooth oscillator circuit, the range of resistance values that can be used is limited. The charging resistor value can't be too low or the UJT will sustain conduction at the valley current after it has fired once and won't oscillate. The charging-resistor value can't be too high or leakage will cause calibration nonlinearity and, in the extreme case, oscillations will cease because the voltage across the capacitor won't build up to the UJT peak point. In the present case it's desirable to use fairly large values for the charging capacitor to eliminate other problems, and at the same time fairly high frequencies are required; thus a low value of charging resistor is necessary, and the ordinary circuit simply will not operate.

This problem was solved by adding an inexpensive high-current transistor to help discharge the reference capacitor. In **fig. 2** the UJT is Q2, and Q1 has been added to furnish a more rapid and complete discharge of C1 or C2. The firing

C1	0.1 μF 50 V Mylar (CD WMF05P1)
C2	1.0 μF 50 V Mylar (CD WMF05W1)
C3	0.005 μF 5% Polystyrene (CRL CPR-5000)
C4	250 μF 25 V (Mallory MTV250DN25)
C5	250 μF 10 V (Mallory MTV250CP10)
C6	700 μF 3V (Mallory MTV700DJ3)
C7	1000 μF 50 V
C8	2000 μF 50 V
CR1	GE Z4LX6.2 1-watt zener 6.2 V
CR2	GE Z4LX9.1 1-watt zener 9.1 V
CR3	GE Z4LX6.2 1-watt zener 6.2 V
CR4	(4) Motorola 1N4001
M1	0-1 mA meter (Simpson Model 524)
Q1	TI 2N5451 (800 mA)
Q2	GE D5E43 unijunction
Q3	TI 2N5451 (800 mA)
Q4	RCA CA3046 5-transistor IC array
Q5	RCA CA3046 5-transistor IC array
S1	Range switch, 2 pole 6-position (Mallory 3236J)
S2	Toggle switch
T1	24 V filament transformer

point of Q2 determines the period of the oscillation for a given RC time constant and fixed supply voltage. As soon as Q2 triggers it furnishes base drive to Q1, which then discharges C1 or C2 with a peak discharge current of around 400 mA. Either capacitor is discharged in a few microseconds to a voltage below the valley voltage of Q2, so Q2 is forced to

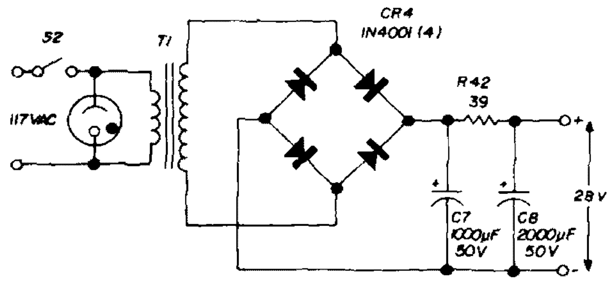


fig. 3. The capacitance meter power supply.

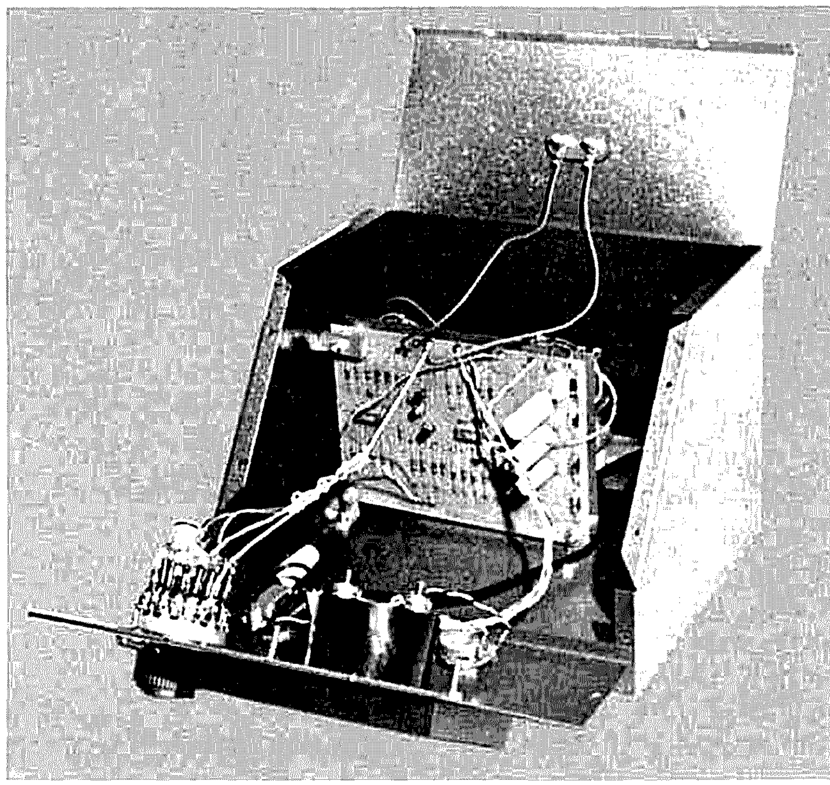
cease conducting despite the low value of charging resistance.

The effect of leakage is also reduced in the oscillator circuit by returning Base 2 of Q2 to a regulated voltage less than the charging voltage. This lowers the UJT peak point; and while the output amplitude is reduced and the waveform has a dc level above ground, the sawtooth waveform is much more linear. Hence the effects of leakage shunting C1 or C2 are greatly reduced.

circuit analysis

The sawtooth generator output appears across the emitter load of emitter follower Q4-3. Voltage divider R13,R14 sets the firing point in the reference sawtooth waveform at which Schmitt trigger Q4-1, Q4-2 changes state. When the voltage at pin 2 of Q4 (junction between R13 and R14) is low Q4-2 conducts, Q4-4 is cut off, and Q3, the discharge transistor for C_X, conducts. The voltage across C_X is thus clamped during this time to the $V_{ce(sat)}$ of Q3. Part way up on the reference sawtooth, the voltage at pin 2 of Q4 rises to the trigger point, whereupon Q4-2 swings to cutoff, Q4-4 conducts, and Q3 is cut off. C_X begins to charge at this instant since it is no longer clamped by Q3.

As C_X charges toward the regulated power supply voltage through a resistance determined by the range switch, a point is reached where the voltage of C_X is high enough to trigger the second Schmitt trigger, Q5-1,Q5-2. At this instant Q5-2 cuts off, the voltage applied to the NAND gate again becomes positive, and Q5-5 saturates. Since Q5-5 and Q4-5 share the same collector resistor, R34, it is apparent that both these transistors must be simultaneously cut off (base negative) for any voltage to appear at the common collector junction with R34. So only during the time when C_X is charging and Schmitt trigger Q5-1,Q5-2 has *not* triggered, is this condition met. Hence the positive pulse that appears at the common collectors of Q5-5 and Q4-5 has a width exactly proportional to the value of C_X. This voltage, the gate output, energizes the meter through emitter follower Q5-4. Capacitor C6 provides additional averaging and was chosen for the lowest frequency range (highest capacitance of C_X) to prevent meter flutter.



Inside the direct-reading capacitance meter. The printed-circuit board, with or without semiconductors mounted and soldered in place, is available from the author. Choose value of R42 for 28-volt output.

calibration

Potentiometer R40 is the calibration control, which is set so that the meter reads a value corresponding to a known capacitance applied to the C_X terminals. The calibration circuit is R10,C3, which may be switched in at position 6 of the range selector. C3 was chosen to give a mid-scale reading on the 0-0.01 μF range (range 2). The unit must be calibrated with no capacitance connected across the C_X terminals. C3 is a 5-percent capacitor; but if a 1-percent capacitor is used, or one whose value has been determined by an accurate bridge, the over-all accuracy of the meter will be improved.

range switching

The range switch uses resistors rather than capacitors, except for the lowest range, because 1-percent resistors are less expensive than 1-percent capacitors. C1 and C2 are the only capacitors switched in the ranging circuit. Their exact value isn't important (e.g., their tolerance can be 10 or 20 percent; but C1,C2 must be reasonably stable with temperature and time and have low leakage). Their ratio should be 10:1. Since C1,C2 will not, in general, have this ratio, it's necessary to trim R1 with an additional series or shunt

resistor so that a known 0.001-μF capacitor, say, when set to read 0.1 mA on M1 when on range 2 will read 1.0 mA when on range 1. In the constructed unit, R1 was trimmed by shunting a 12k ½ watt 10 percent resistor across it.

leaky capacitors

These components will show a higher apparent capacitance than their true value. However, the leakage has to be pretty bad for this to occur because the maximum charging resistance in the C_X circuit was kept deliberately low – a maximum of 20k on the lowest capacitance range to 1k on the highest range. If the capacitor to be measured shows less than about 10 times the charging resistance for the range in use, when measured with a good ohmmeter, some error will occur in the capacitance measurement. This much leakage would be unacceptable in most circuits anyway, so the capacitor should be discarded.

The five ranges are (in μF):

Range 1	0-0.001
Range 2	0-0.01
Range 3	0-0.1
Range 4	0-1.0
Range 5	0-10.0

In Range 1 the lowest value of capacitance that can be accurately measured is about 100 pF, or about 10 percent of full scale on the meter. Stray capacitance in the leads to the selector switch, etc., will cause the meter to indicate slightly on Range 1 with no capacitance connected to the C_X terminals. Also on this range, the reference frequency sawtooth is fairly high; and even though the discharge time of C1 is only a few microseconds, it is not an altogether negligible percentage of the sawtooth period. This, together with other effects such as finite rise and fall times in the waveforms, limits the minimum capacitance that can be accurately measured.

power supply

The power supply schematic shown is used in the unit pictured, but it can be any unregulated 28-volt supply delivering 50 mA with reasonably low ripple. Zener diodes CR1-CR3, inclusive, compensate for line voltage variations to which the instrument would otherwise be sensitive.

*Those interested mav obtain the printed circuit with or without all semiconductors installed. Write the author for details.

construction

The direct-reading capacitor meter is mounted in a Bud C1585HG sloping panel cabinet. The unknown-capacitance terminals are a pair of binding posts on top. The main circuit is on a 4¼ x 6 inch, G-10 epoxy etched circuit, with two-sided etching and plate-through holes.* The power supply is mounted on a piece of printed circuit material the same size as the main circuit, and bolted to it by spacers. The assembly is then secured inside the Bud case by right-angle brackets.

conclusion

This instrument has been useful in checking unknown capacitors from the junk box and from surplus equipment. A surplus audio filter tuned to 1050 Hz, for example, contained many unmarked capacitors that had to be measured to decrease the filter frequency to around 400 Hz for more comfortable cw listening. Dried-up electrolytics, supposedly 10 or 8 μF, were found to be less than 1 μF when measured. Some ceramics read as much as ten times their GMV value. It was also interesting to study the temperature drift of various capacitor types by watching the meter move as a soldering iron was held near one lead of the capacitor or as cold chemical spray was applied.

Finally, if you need a more precise measurement than is possible with the Simpson 524 (i.e., when matching two capacitors for an equal value or a given ratio), an oscilloscope can be used to measure the duration of the NAND gate pulse appearing across R39 in the circuit. If you have a burning desire for a digital readout, the NAND gate pulse width can be read with a time-interval meter and Nixie or LED displays.

ham radio

oscilloscope voltage calibrator

This simple voltage calibrator increases the utility of your scope with calibrated voltage-reference points

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As **anybody** who **regularly uses one** will agree, the oscilloscope is one of the most versatile electronic instruments available and in many situations is indispensable. A scope may be a relatively simple general-purpose device or a greatly refined and complicated, high-precision, laboratory instrument. Its cost may run from considerably less than one hundred dollars to many hundreds of dollars. Its most important use, of course, lies in the study of alternating-current waveshapes; whether such waveshapes are regular in form, as in shine, sawtooth, square, trapezoid or pulse-train shapes; or irregular, as in speech, clipping, video, integrated sync pulses, transients or other distorted-type forms.

In addition to actual waveshape observation, a very useful application of the scope is the measurement of the voltages of waveshapes for which averaging-type meters, such as multimeters and vtvms, are unsuitable. This usage requires that the scope be calibrated as a voltmeter; this article describes an accessory which will make less-expensive scopes more useful in this respect.

There are many thousands of general-purpose scopes in use. Most of them do not have a voltage calibrator as an integral part of the instrument. If a calibrator is included it consists of a single calibration

voltage of doubtful accuracy. By the use of the simple accessory described here, the utility and voltage measuring accuracy of these scopes may be greatly enhanced. An accurate square wave is ideal for calibration usage. This accessory develops such a waveform.

circuit

The schematic, fig. 1, is straight forward. Parts are easily obtainable and no trick circuitry is involved. It works beautifully and can be assembled in one evening's time. Parts placement is in no way critical. A shielded box, such as a small LMB box, not only makes a nice package, it will also prevent noise pickup from reaching the scope. The switched outputs give square-wave voltages of 1, 2, 3, 4 and 5 volts peak-to-peak.

The rectifier diode CR1 is specified at 1 amp so that it will handle the charging current for the electrolytic capacitor. The zener diode assures a constant and stable dc reference voltage for fluctuating line voltages. Nearly any silicon diodes may be used for CR2, CR3 and CR4 but all three should be the same type. The circuit current is normally less than one mA.

Resistors R1, R2 and R3 may be 10% values (silver band). Resistors R5, R6,

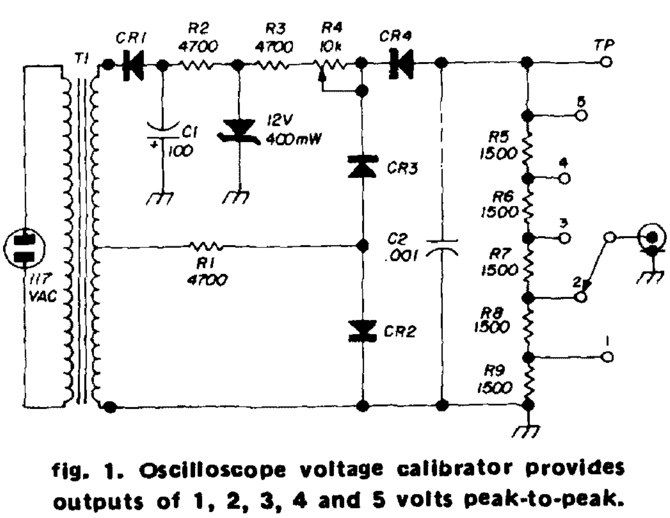


fig. 1. Oscilloscope voltage calibrator provides outputs of 1, 2, 3, 4 and 5 volts peak-to-peak. Resistors R5, R6, R7, R8 and R9 should have 5% tolerance. Diode CR1 is 100 PIV, 1 A. Diodes CR2, CR3, and CR4 are silicon diodes with 1-mA rating (1N4005, 1N914, etc.).

R7, R8 and R9 do not have to be precision resistors. Ordinary, good quality ½- or ¼-watt carbon composition resistors, 5% tolerance (gold band), will do very nicely for the voltage divider. The potentiometer, R4, should preferably have a linear taper although an audio taper is usable.

Because a stable dc reference is used, this potentiometer may be mounted inside the box since once it is set it should require no further attention. The switch should be a non-shorting type. Suit your personal preference regarding the output jack, but since the lead from the calibrator to the scope should be shielded, a BNC or good phono type should be used. The use of shielded lead (preferably coax) will prevent hum modulation of the calibrator's square-wave output.

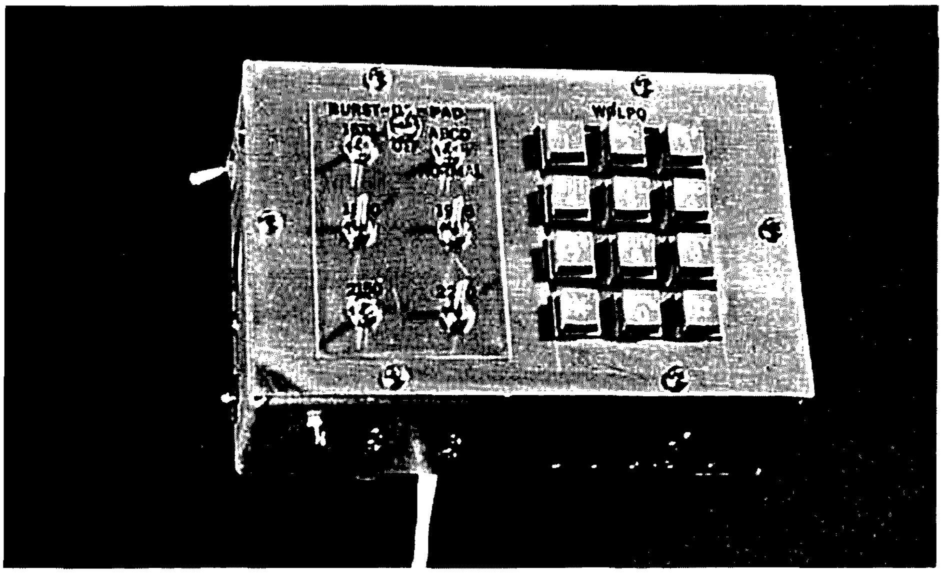
calibration

The test point TP is used when adjusting the calibrator output voltages. Use a good vtvm on its 3- or 5-volt dc range (positive to chassis, negative to TP). Adjust R4 to give -2.65 volts dc, as accurately as you can, at the test point. Your calibrator will then be calibrated for a 5-volt peak-to-peak square wave. Switch positions will provide one-volt steps; thus, peak-to-peak outputs will be 1.0, 2.0, 3.0, 4.0 and 5.0 volts.

On dc-coupled scopes there will be a perfectly flat top and bottom to the wave. On ac-coupled scopes you will notice a very slight slant to this line; on such scopes a reduction of the horizontal gain will reduce its effect on the visual accuracy with which you can read the calibrated lines on the scope.

With the usual X1, X10 and X100 step attenuator positions on most scopes, your scope gain control will permit you to set a wide range of accurately known voltage values for any scope division you may wish; such as 1.0 volt, 0.1 volt or 0.01 volt, or perhaps 4.0 volts, 0.4 volt or 0.04 volt per centimeter or per inch or per any other division marking you may desire.

ham radio



mobile operation with the touch-tone pad

Making a clean and dependable mobile fm installation with a touch-tone pad

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Numerous articles have been written concerning the utilization of Touch-Tone for control functions on vhf and uhf repeater systems, even going so far as to include partial control of an existing hf system via the uhf radio link. However, articles that deal with actual construction and operation of these Touch-Tone pads are few.[1,2] This article hopes to fill the void, and help some enterprising amateurs build operational Touch-Tone systems.

The Touch-Tone pad enclosure shown in fig. 1 has been in use for over a year with great success and much satisfaction. My unit is being used with the Regency HR-2. Both the HR-2 and the HR-2A are quite popular in this area, and to date, no one has experienced any difficulty with this circuit. Over the last year there have been some basic design changes, most of which are incorporated in this article.

the pad

The Touch-Tone pad I used is the Western Electric Model 35A3. I understand the Model 35Y3 is basically the same, only a newer model. The schematic diagram shows color coding of each wire on the 35A3 and to what point each wire

goes. Note that the *red* lead is not used, and is taped or clipped off. The *green* lead is the power and audio take-off point. There are two leads left that are rather unusual. They are the *white* and the *white and blue* leads. Both leads terminate in the pad at a switch contact which is shunted by a 5.1k resistance. When *any* button on the pad is depressed, the switch contacts open giving this 5.1k dc resistance. Both leads are free from any *other* contact or connection within the pad, leaving them free for control of an external circuit, which, in our case, will be push-to-talk keying of the vhf transceiver.

operation

The keying circuit: The two leads (*white* and *white and blue*) from the pad are fed to a transistor switch. One lead (*white*) is grounded, and the other (*white and blue*) is tied to the base of Q1 (2N2222A, 2N718A or other npn). With the pad in an idle condition, the base of Q1 is grounded and is cut off. When any button on the pad is depressed, the base is returned to ground through the 5.1k resistance in the pad. This resistance plus the external 2.2k resistor form a voltage divider, forward biasing the transistor switch, Q1.

The relay: The relay is a subminiature dpdt type with a holding current of about 30 mA. Both sets of contacts are used, one for push-to-talk keying of the vhf transceiver, and the other to toggle the microphone audio of the transceiver between the microphone and the output of the Touch-Tone pad. Breaking the microphone connection insures that no background audio can reach the transceiver microphone input causing distortion of the Touch-Tone tones.

Across the relay coil is a series combination of R1 and C1 and parallel diode CR1. The capacitor C1 sets a delay (relay dropout) time of about 1.5 seconds, allowing the user time to complete a dial sequence *without* the tranceiver dropping out of key. This feature prevents excessive wear of the transceiver relay contacts. It also allows completion without the chance of missing any tones, in case you are a fast dialer.

Dc power source: If you plan to utilize this unit with an existing base station supply, then no problems should be encountered. But, if you are going mobile, the combination of zener diode CR2 and the 33-ohm resistor should be included. This unit likes a stiff dc source.

If your particular installation is noisy, it may be wise to include more than 1500 mF of filtering across the zener. I have just purchased a very noisy car. I had to add an additional 1500 mF of filtering, which has apparently cleaned up the system. Installation of a coaxial bypass capacitor at the firewall helps the filtering. All cables should be shielded, even from the battery to the firewall bypass capacitor. You can even add an inductor in series with the coaxial bypass, with as much as 2000 mF to ground for that extra measure of protection. It has been found that it is hard to have too much filtering in the automotive system.

Pad impedance and levels: The Touch-Tone pad is a *low-impedance unit*. Variable resistor R2 should be not more than 1k. The specified value of 500 ohms is preferable, otherwise you will have to be almost at the bottom of the pot to keep the tones from distorting. The electrolytic capacitors (1 mF and 2 mF) feeding the audio output line are there to preserve the low-frequency characteristics of the Touch-Tone pad. The 100k resistor, R3, is included to isolate the microphone input from the pad, and to present a high impedance for the microphone input of the transceiver.

Note that there is also included a 1M resistor from the output of the tone burst generator. This value may require some experimentation by the user to obtain the required deviation of the transceiver. If your tone burst generator is one of the types with a very high output level, this resistor may be a value of several megohms. If your transceiver is equipped with *any* kind of clipping, then pot R2 should be set for a level that is below clipping. If it is not, then the receiver Touch-Tone decoder will not recognize

any of the tones, as they will not be true sine waves. These values (1M) work well with the Regency HR-2. If your radio is a different brand, you may need to juggle them to achieve the desired results. But, in no way should these resistors be less than 47k.

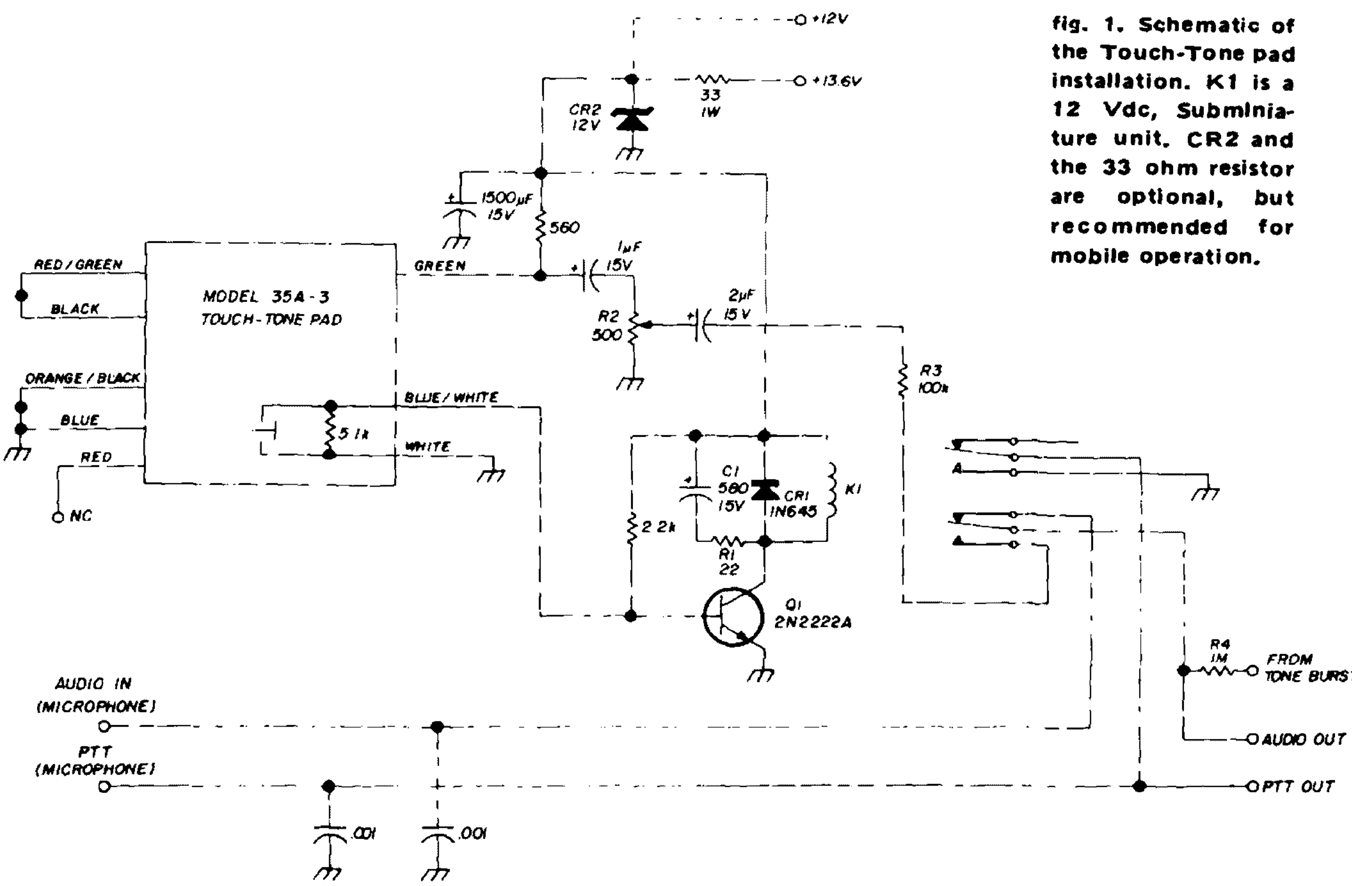


fig. 1. Schematic of the Touch-Tone pad installation. K1 is a 12 Vdc, Subminiature unit. CR2 and the 33 ohm resistor are optional, but recommended for mobile operation.

construction

My chassis box is 4 x 6 x 2 inches. The pad is mounted on the right side of the box, and is held in place by two 6-32 screws. One screw mounts on the right hand wall of the box, and the other mounts via an angle bracket secured to the bottom of the box. The tone burst generator is a small pc board (1½ x 2½ inches) mounted on the left wall of the box. I have used a matching microphone jack, which accepts the HR-2 microphone. Shielded Teflon cable runs to the HR-2. In my case, the Teflon cables are soldered directly into the enclosure, and routed out via a small hole in the back of the box. However, phono connectors would make it easier to connect and disconnect cables if you ever wanted to work on or modify your pad.

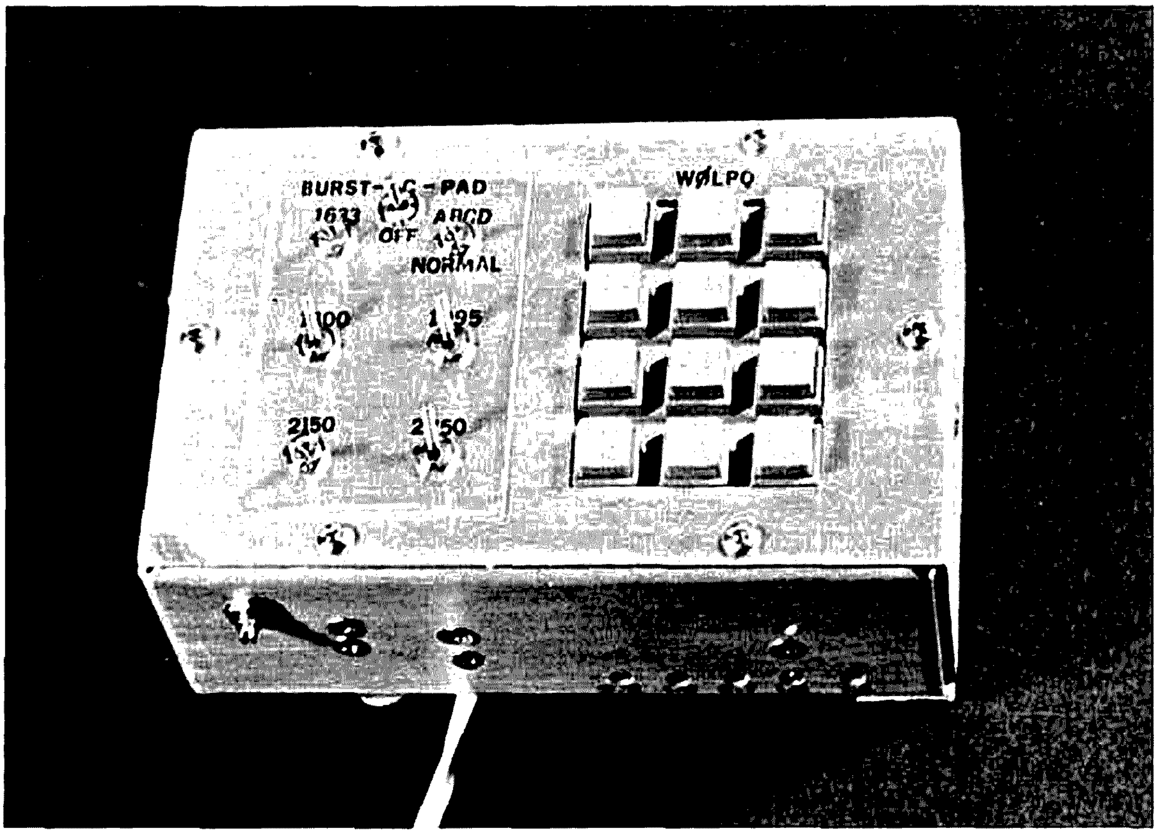
Cutting holes for the pad is a matter of personal choice. You can either make one large hole to show all of the buttons, or get more ambitious and cut individual holes for each button. If your choice is the latter, then using an existing Touch-Tone instrument (assuming you have Touch-Tone in your home) for exact placement of the holes, will leave you with a very neat enclosure. A nibbling tool is very handy for this, but do it from the inside of the box, or your end result will be a disaster area for a front panel.

In the photograph of my unit, I have used individual toggle switches for the tone burst frequency selection. A rotary switch *could have* been used, but a suitable subminiature type was not available. The center toggle switch which is labled BURST-DC-PAD, selects dc power

for either the tone burst generator or Touch-Tone pad. My feeling is that I will never require power to both at the same time, so why not toggle the power between them? Also, why have a beep on the air if you are not using a repeater access channel? A clean radio installation goes a long way toward making friends, and in this age of radiomania, who needs enemies? We've got enough now, why add to the collection?

trouble

Murphy and his laws always get into the act. Again, I caution the reader about the +12 V supply line. Why? Well, reread the last two sentences of the preceding paragraph. If you could hear some of the mobile installations around here, you would not ask why. Alternator whine abounds inside that engine compartment. Even with the installation of resistor plugs, radio ignition wiring, and the rest of today's modern innovations, it is still there. But, to keep peace with the neighborhood ham population, take a little patience, and do it right the first time. You may need to add some more filtering, but really, isn't it worth it? If you are plagued by stray rf floating around your mobile, then add a couple of 0.001-mF capacitors across the microphone audio and push-to-talk lines inside the pad enclosure.



The pad and associated circuitry are mounted with a toneburst generator in one box.

Special thanks are due to Tom Yocom, WAØZHT, whose digital mind dreamed up the repeater with which this pad is used. Also, thanks are due to Jerry Buck, KØYBM,who designed the tone burst generator used in this system and who also did the photograph for this article.

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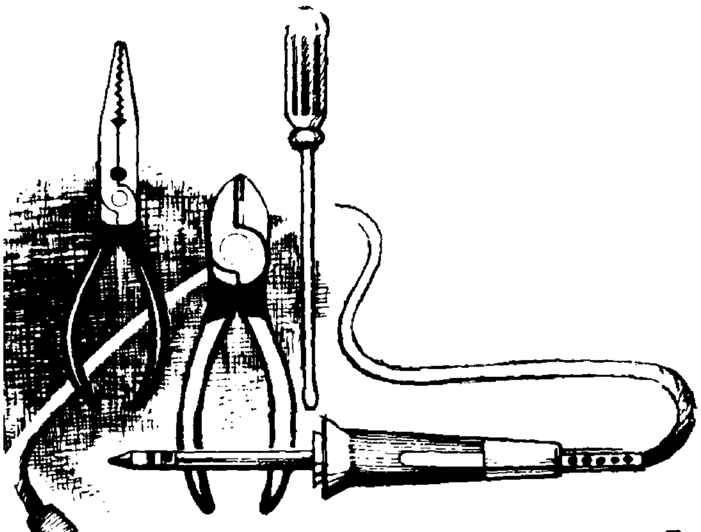


circuits and techniques

ed noll, W3FQJ

digital IC oscillators and dividers

Modern counters and calibrators feature crystal-controlled multivibrators using NAND and NOR gates plus digital IC divider chains. A popular low cost unit is the 7490 decade divider. In its in-line case it houses separate 5-to-1 and 2-to-1 counters, fig. 1. The 5-to-1 counter is actually an inhibited group of three individual 2-to-1 units as described in detail in last month's column. The 5-to-1 and 2-to-1 combinations have separate input and outputs. The output of one is wired externally to the input of the other to obtain an overall count of 10-to-1. This wiring can be done in two ways; the output of the 2-to-1 fed to the input of the 5-to-1, or, depending upon the desired intermediate count, the output of the 5-to-1 applied to the input of the 2-to-1. The device has great versatility in terms of the selection of preferred counts.

Let's consider these possibilities in terms of calibration frequencies, assuming a 100-kHz drive signal and two decade dividers. Note in fig. 2A that calibration points would be made available at 100, 20, 10, 2 and 1 kHz. If you take the first decade counter and connect the 2 to the 5 rather than the 5 to the 2, a different combination is set up consisting of 100-, 50-, 25- and 2.5-kHz points.

Additional combinations are shown in figs. 2C and 2D. There are other possibilities, too, that can be established using interwiring combinations as shown in fig. 3.

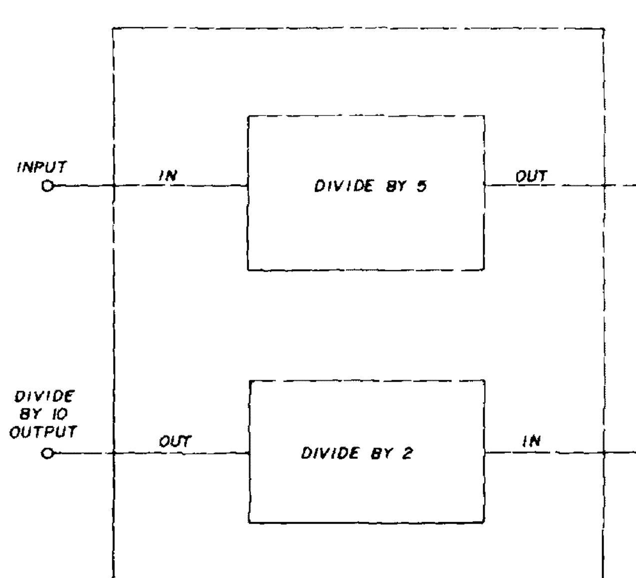


fig. 1. Basic plan of the 7490 decade divider.

All of the above is quite understandable. However, to the uninitiated, the wiring of an integrated circuit appears to be a very complicated thing. Actually it is simple and the major complication is usually the printed-circuit board. However, this can be avoided by using straight wiring techniques as suggested in the first experimental procedures in the June column. If you use binding posts and jumpers it is also possible to change the count sequences between the combinations shown in figs. 2 and 3.

The pin-out wiring diagrams for the 7490 are given in **figs. 4** and **5**. Note how very simple it is. There are a number of terminals to which no connection is made and another group which are all tied to common. Of course, there are supply voltage as well as input and output connections to be made. The diagrams of **fig. 4** are for using the 2-to-1 and 5-to-1 counters separately. Both the connections of fig. 5 provide the 10-to-1 count.

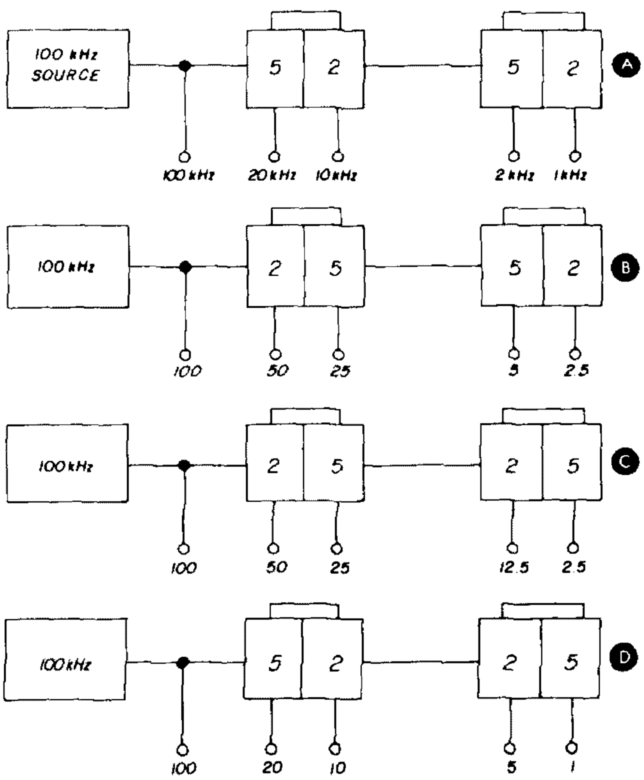


fig. 2. Some of the count possibilities using two decade dividers.

However, in the first example, the first count is 5-to-1; the second, 2-to-1. The second example is the converse, using the initial 2-to-1 count and then the 5-to-1.

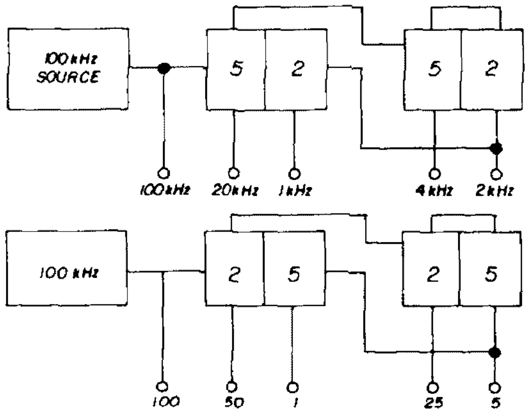


fig. 3. Two additional count possibilities using two decade dividers.

digital IC oscillators

Digital ICs of suitable design can also be used as high-frequency crystal-controlled square-wave generators. The 7400 NAND gate used initially in this series can be operated as a high-frequency oscillator. Two of the four gates are wired as a multivibrator while a third one is used as a buffer output. Doug Blakeslee, W1KLK, has used this common IC successfully with the circuit of fig. 6A.[1] Its output is followed by two 7490 decade dividers.

Ted Bensinger, W5PCX, uses the 7400 in the 3-MHz IC oscillator arrangement of fig. 6B.[2] Two of the NAND gates again serve as the multivibrator while the two other sections are pressed into service as buffer and calibrate outputs. Two decade dividers provide the countdown to 30 kHz. W1KLK operates his circuit at 3 MHz to get the same 30-kHz output. However, he employs a high-frequency 74H00 NAND gate. Theoretically this IC should provide steeper sides and higher harmonic output levels.

quency is wise because a division of 100 results in a 30-kHz output which is the standard channel separation on the two-meter fm band. The 3-MHz crystal frequency can be adjusted and set on its fifth harmonic which falls on the 15-MHz WWV frequency.

An elaboration of the W1KLK calibrator could include a 3-to-1 counter providing a 1-MHz output, fig. 7. In so doing, a calibration point would be available on other WWV frequencies, if and when the 15-MHz signal is not receivable in a given area. Furthermore, the 1-MHz output could also be switched into the divider chain to obtain the same sequence of output frequencies made available from a 1-MHz crystal source.

The dual binary 7473 J-K flip-flop used in last month's experiments can be wired as a 3-to-1 counter as shown in fig. 8. Typical waveforms are shown in fig. 9. Another 1-MHz circuit using an MRTL NOR gate is shown in fig. 10.[3]

higher-frequency operation

High-speed digital ICs permit crystal-controlled square-wave generation

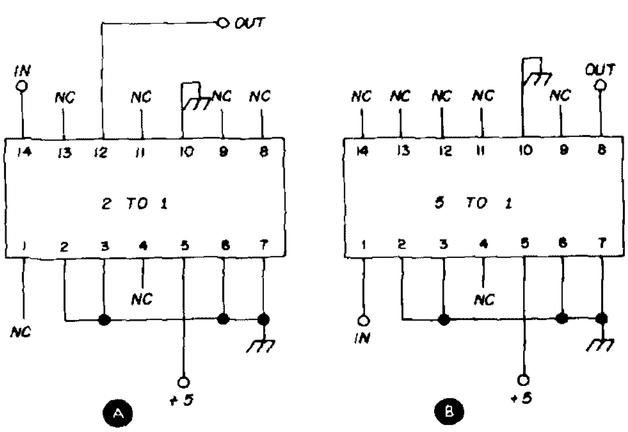


fig. 4. Pin-out diagrams for wiring 7490 decade dividers.

at even higher frequencies. The advantage of an output square wave, of course, is in its high harmonic content. A 5-MHz oscillator of this type produces strong discernible calibration points on the amateur vhf and uhf bands. Typical circuits for an MECL type are given in fig. 11.[4]

Circuit A is for fundamental crystals over a frequency range from approximately 1 MHz to 20 MHz. Circuit B has a broadly resonant output circuit and works with

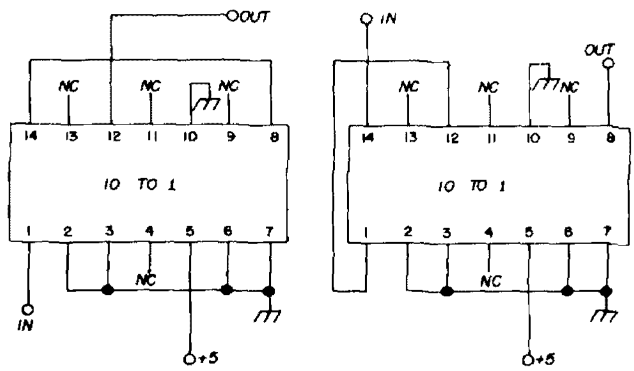


fig. 5. Dividing by ten in 5 x 2 or 2 x 5 sequence.

overtone crystals. I have used this latter circuit successfully with overtone crystals in the 50- to75-MHz range to produce marks on the 6-, 2-, 1¼- and 3/4-meter bands. Using a uhf television receiver, calibration points have been checked all the way through the uhf band.

The vhf oscillator can be checked out by using a double calibration technique. I calibrate a 5-MHz crystal oscillator on the highest receivable WWV frequency at the moment. I then tune in its tenth harmonic at 50 MHz on my six-meter receiver. A 50-MHz overtone crystal in the circuit can then be zero-beat with this tenth harmonic of the 5-MHz crystal. Numerous other possibilities exist.

meteor scatter

I have written before of R.A. Ham and his studies of vhf propagation. Mr. Ham is an amateur radio astronomer. Although I have never met or corresponded with him, I have an exceptional respect for his work because of his skill and his individual interest in basic research. He is interested in a phenomenon because it is presently relevant or practical. Our sently relevant or practical. Our very obsession with "now" relevance and our "so-what" attitudes have become

self-destructive to us as individuals and as a society. One must realize that those things which are non-relevant today might well be of dire relevancy a century from now or even a decade from now; and, furthermore, what we consider relevant today might be most insignificant tomorrow.

Again the following meteor observations of Mr. Ham demonstrate how important radiobeacons can be in basic propagation research. The following is a quote from his coverage in *Electronics Weekly* (British):[5]

"The earth, during its annual orbit of the sun, periodically encounters large swarms of meteor particles which are known to astronomers as meteor showers. Reference to the annual handbook of the British Astronomical Association will show that the major meteor showers are the Quadrantids in January, the Lyrids in April, the Perseids in August, the Leonids in November and the Geminids in December.

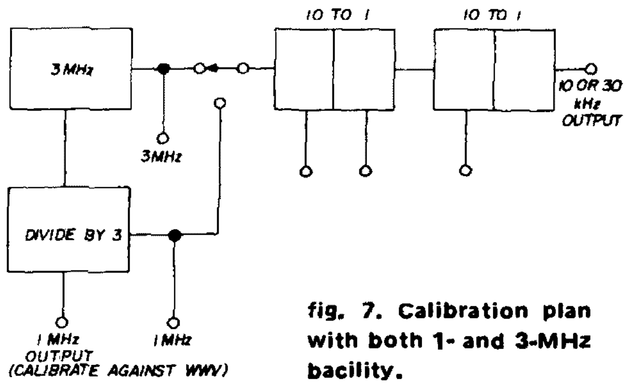


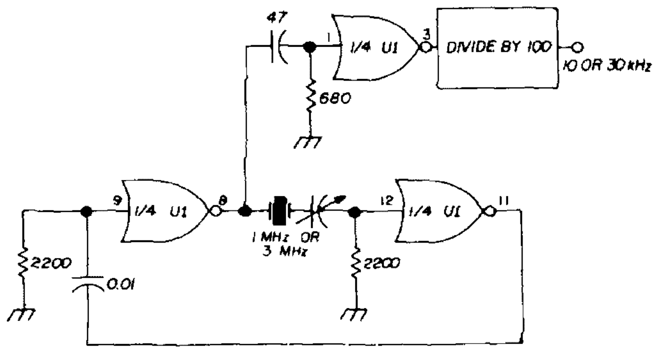
fig. 7. Calibration plan with both 1- and 3-MHz bacility.

"The name given to each meteor shower is derived from the constellation of stars from which direction the radiant of the meteors appears to come. For example, one would look toward the constellation of Perseus for the radiants of the Perseid meteors, and toward Leo for the Leonids.

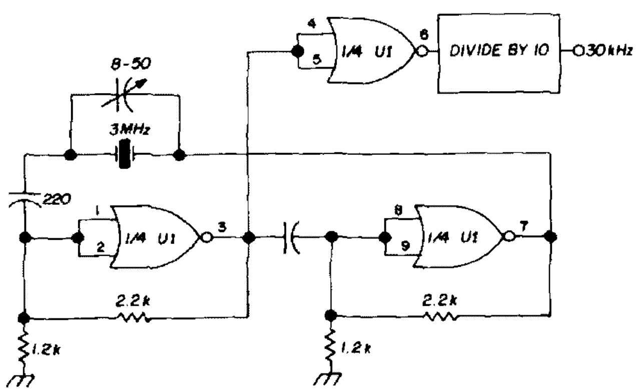
"The *visual astronomer* hopes for clear skies to enable him to make meteor observations and estimate the number of meteors within the area of the sky which he can see. It is common practice for groups of observers to combine their efforts during a meteor shower to gain as much information as possible. Their reports describe the color, direction and duration of the meteors which they have seen.

"Unfortunately, the visual astronomer has two enemies to contend with; the moonlight, which can make the sky too bright for satisfactory observation, and the weather, which can cloud over the sky at the vital time and prevent any observation at all.

"The amateur radio astronomer can assist his visual colleagues without being hampered by moonlight or overcast skies. The dying ionized trail left by a meteor provides a short-lived passive reflector which can bounce radio signals over a



A



B

fig. 6. Using the 7400 as a high-frequency oscillator. Use a 7400 with a 1-MHz crystal and a 74H00 with a 3-MHz crystal. Both the 7400 and the 74H00 are quadruple NAND gates, but only three of the gates are used in these oscillators.

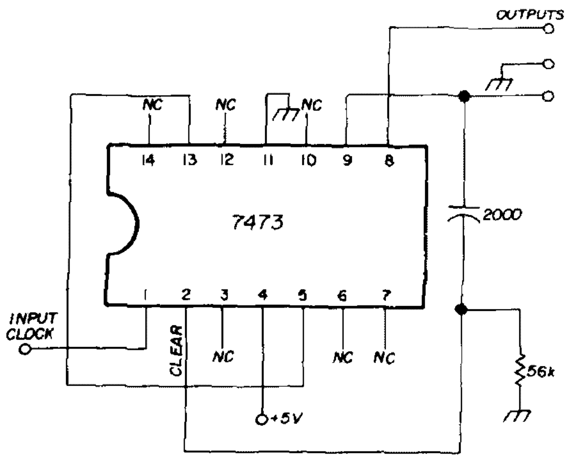


fig. 8. The 7473 3-to-1 divider using two binary flip-flops.

thousand miles. A radio receiving station can tune to the frequency of a distant transmitter and record the number of times that its radio signal is deflected by ionized meteor *trails*.

"During the life of a meteor shower, there can be a large amount of temporary ionization within the earth's atmosphere, and several United Kingdom radio amateurs have taken advantage of this and established communication, on the vhf bands, over large distances. This means of communication is known as the meteor scatter technique.

"Many amateur radio operators within the past two decades have established two-way meteor-trail communication with fellow amateur stations a thousand miles away. These complicated contacts are confirmed by the exchange of QSL cards between the two stations concerned.

"Attempts to communicate via meteor scatter are usually at pre-arranged times

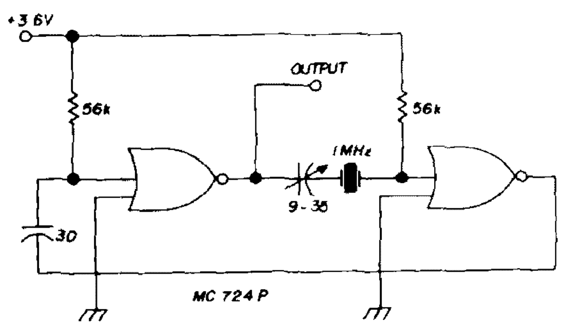


fig. 10. MRTL NOR gate as a 1-MHz oscillator.

and are conducted during a meteor shower when there is a high chance of success.

"Bearing in mind that each meteor trail may only survive for a few seconds, each operator involved in the contact must repeat his call sign and message in high-speed Morse code many times. This *intermittent* repetition enables the receiving end to piece together tiny bits of the transmitted signal.

"Both operators may have to repeat this procedure at regular intervals for over an hour before a two-way contact can be confirmed.

"Using their CW *skill* and extreme patience, several United Kingdom ama-

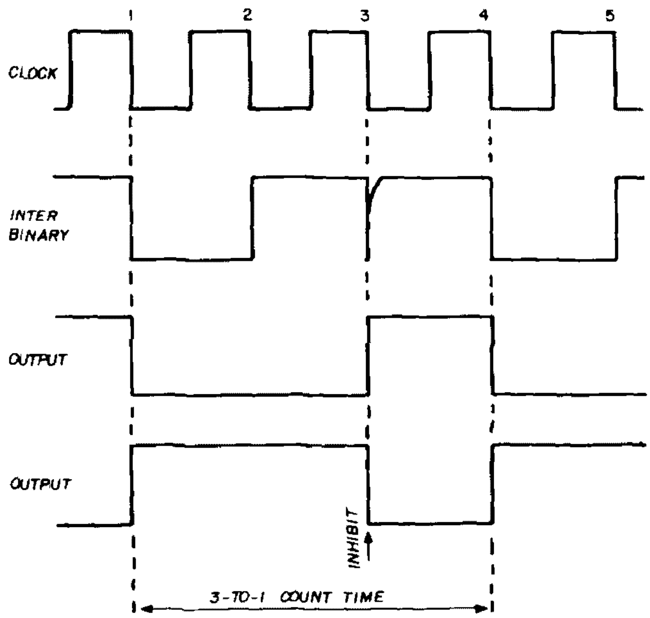


fig. 9. Three-to-one count waveforms.

teurs have confirmed meteor-trail contact with amateur stations in some twenty countries. The majority of this work is with continental amateurs and carried out at the CW end of the two-meter amateur band (144.0 to 144.1 MHz).

"Should a reader hear an amateur during a meteor shower, repeating his own call sign at high speed, take a listen on his frequency when he has finished transmitting and minute parts of the replay may be heard.

"Using meteor trail reflection for communication is a chancy business, but I

personally think that it is a sporting method and a great achievement.

"During the 1968 Geminids, the author (at his Sussex home) conducted an experiment using the RSGB beacon at Thurso in Scotland as the transmitting end. The object of the experiment was to produced 'GB3' and over 100 minutes later at 2205, another large trail enabled '3GM' to be recorded. The radio signal from a beacon is a continuous tone, which is frequently interrupted with the beacons identification sent in Morse code."

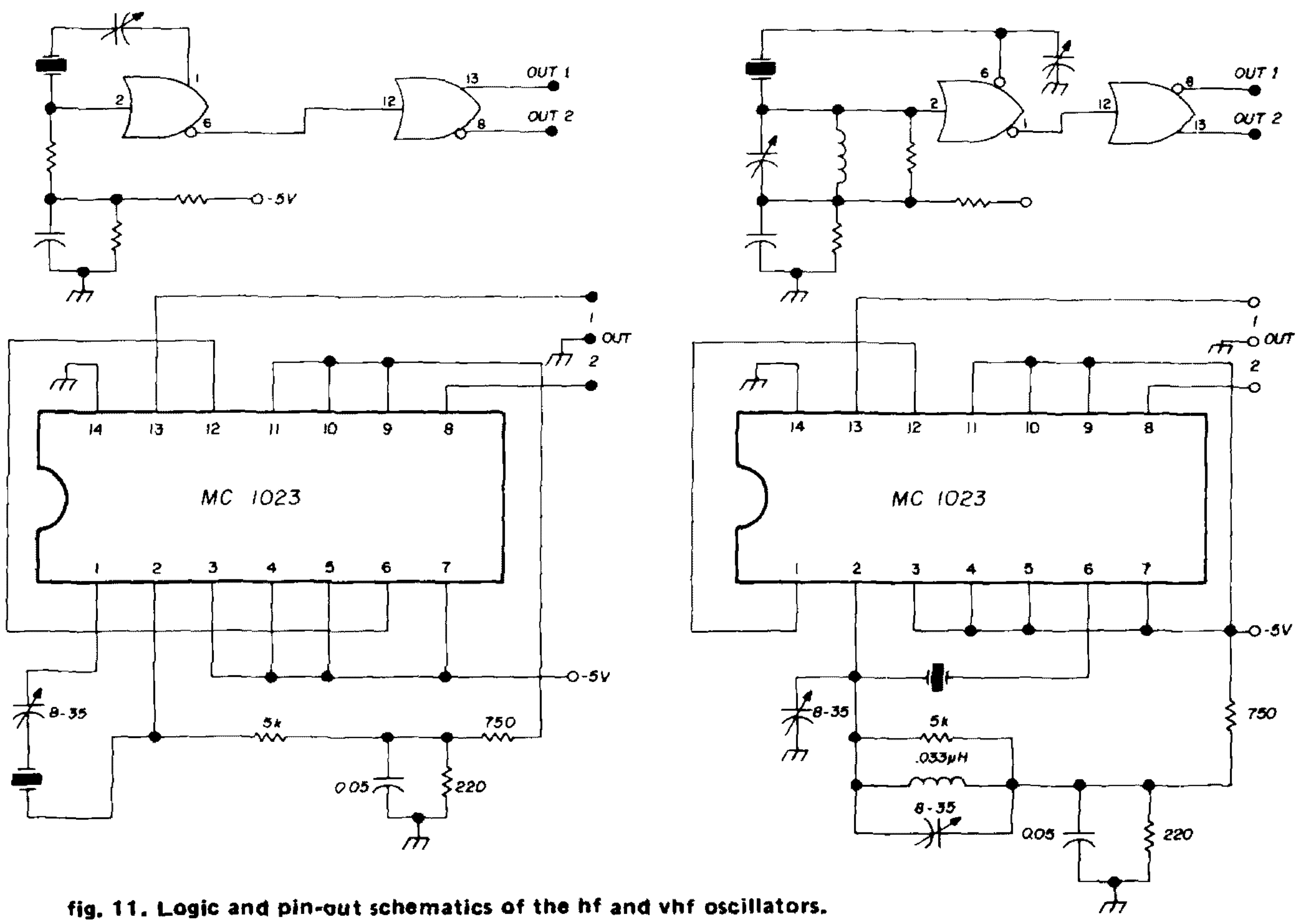


fig. 11. Logic and pin-out schematics of the hf and vhf oscillators.

evaluate the possibility of recording the beacon's identification (GB3GM) by the meteor-scatter technique.

"On December 13, 1968 at 2000 GMT, the author directed his four meter beam toward the north, carefully tuned his receiver to the beacon frequency (70.305 MHz), and accompanied by an interested witness, monitored this frequency until 2300 GMT.

"Throughout the three-hour period, 'pings' of the beacon's tone could be heard bouncing off tiny meteor trails. At 2049 a deflection from a larger trail

references

1. Douglas A. Blakeslee, W1KLK, "Double Standards," *QST*, April, 1972, page 13.
2. Glen E. Zook, K9STH/5, "F.M. Channel Marker Generator," *CQ*, April, 1972, page 41.
3. Bob Botos, "A Frequency Counter using Motorola RTL Integrated Circuits," Application Note AN-151, Motorola Semiconductors, Phoenix, Arizona.
4. Chuck Byers, "IC Crystal-Controlled Oscillators," Application Note AN-417, Motorola Semiconductors, Phoenix, Arizona.
5. R.A. Ham, "Report on Recent 32-Day Meteor Observation," *Electronics Weekly*, November 3, 1971, page 16.

ham radio

comparison of fm receiver performance

FM equipment manufacturers use two different methods of describing receiver performance — this article discusses the differences between these two systems

Vern Epp, VE7ABK, Box 371, Nelson, British Columbia

Many radio amateurs are confused when they compare the fm sensitivity and performance claims of the various manufacturers. And they have every right to be confused because the methods of measurement are quite different.

Some manufacturers use the 20-dB quieting method, while others use the 12-dB SINAD technique. A comparison of these two very different measurements can be best illustrated with the graph in **fig. 1**. In this graph the zero dB receiver output reference is the output of an unsquelched receiver without an input signal after the volume control has been adjusted so the receiver delivers rated audio output with an fm generator deviating 0.67 at 1000 Hz.

Curve 1 shows noise quieting, and it is plotted by varying the output of an unmodulated rf generator and measuring receiver output. The 20-dB quieting point occurs at approximately 0.45 μV. This is the most simple sensitivity test which can be performed.

In the SINAD test a modulated rf signal generator is used. In this test, receiver output initially increases with increasing signal level (curve 2). The limiters become saturated at approximately 0.5 μV and receiver output levels out at +4.5 dB. The actual audio in the output signal is signal plus noise plus distortion (S + N + D).

Now curve 3 must be plotted, using an rf signal generator which is modulated with 1000 Hz. A distortion analyzer is connected to the output of the receiver and nulled to reject the 1000-Hz signal. Then the remaining noise and distortion can be measured.

Noise and distortion declines quite rapidly as input signal increases. When the input signal is large, the noise component is practically nil, so curve 3 levels out at the inherent distortion level of the receiver.

The difference between curve 3 and curve 2 is the SINAD ratio. The 12-dB SINAD sensitivity is at approximately 0.25 μV input as shown in **fig. 1**. The SINAD ratio is computed from the following formula:

$$\text{SINAD ratio} = (S + N + D) - (N + D)$$

I will not discuss the pros and cons of each of these systems, but both have their place in determining the actual performance of an fm receiver. I can only hope that the individual manufacturers will quote both results so amateurs can make realistic comparisons.

Most amateurs talk about receiver sensitivity in terms of microvolts. The graph in fig. 1 is also calibrated in dB relative to 1 watt, or dBW. In a 50-ohm system, a 0.5 μV signal is equivalent to −143 dBW. (When doubling the rf input

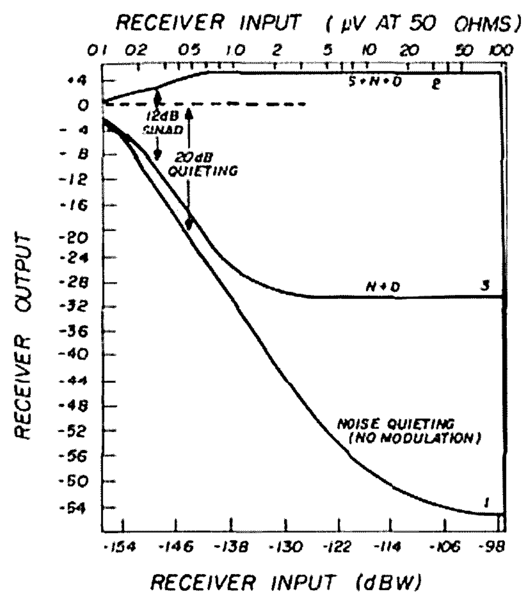


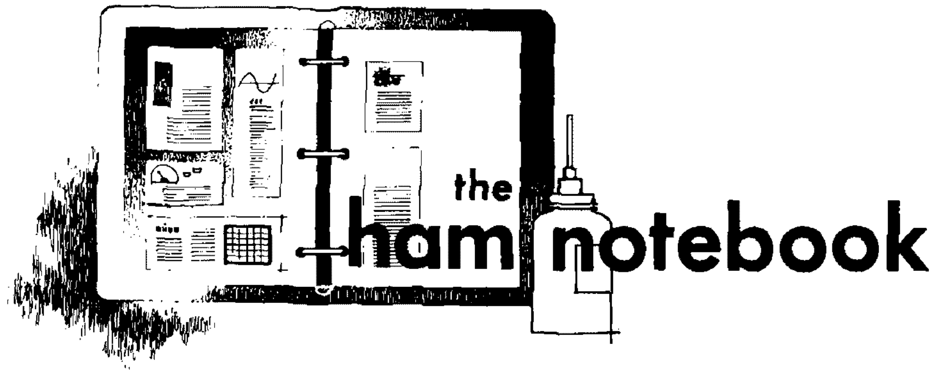
fig. 1. Comparison of fm performance tests. 20-dB quieting is the difference between the zero reference level and noise quieting (curve 1). 12-dB SINAD is the difference between signal plus noise plus distortion (curve 2) and noise plus distortion (curve 3).

signal, remember that you must increase dB by six.) Another term which is used occasionally is dBm, power in dB relative to 1 milliwatt; dBW can be converted to dBm by subtracting 30 dB. For example, −143 dBW = −113 dBm. A complete nomograph of power, impedance and dBm is given in reference 1.

reference

1. W. E. Pfiester, Jr., W2TQK, "Power, Voltage and Impedance Nomograph," *ham radio*, April 1971, page 32.

ham radio



solid-state vibrator replacement

With the flood of solid state vhf rigs on the market, the price of the vibrator powered units has dropped to a level where almost anyone can afford to go vhf mobile. This solid-state switch fits in a small minibox and has only two wires that connect to the vibrator socket (not counting the chassis ground). The only addition to the rig is a small spst switch mounted on the power-supply chassis. Yet this simple unit eliminates the hash and mechanical problems common to vibrators.

The schematic, fig. 1, shows a basic vibrator primary circuit and a secondary winding with buffer. It's basically a mechanical switch with the centertap of the primary at plus 12 Vdc. The vibrator alternately grounds point (A) then (B) creating a square wave of dc. This causes current to flow in the secondary winding. The induced currents always have some transients on their edges. These transients are bypassed by an RC combination called a buffer circuit. This is a resistor and capacitor series combination in parallel with the secondary high-voltage winding.

The schematic in **fig. 2** shows the oscillator. This unit duplicates the electrical function of the vibrator and thus could be likened to a solid-state switch. With power *applied, the* circuit first grounds terminal A then terminal B. A brief description would indicate the toroid (similar in dimensions to an 88-mH toroid) is the heart of the circuit. Its *function is to feed* back current to the bases, alternately saturating Q1 and Q2. CR1 and CR2 are zeners suggested by the transformer company to protect the power transistors from transients. They could be left out but they are inexpensive insurance. The R1, R2 and R3, R4 combinations are resistive *current* dividers, limiting the base current to a saturated but safe level.

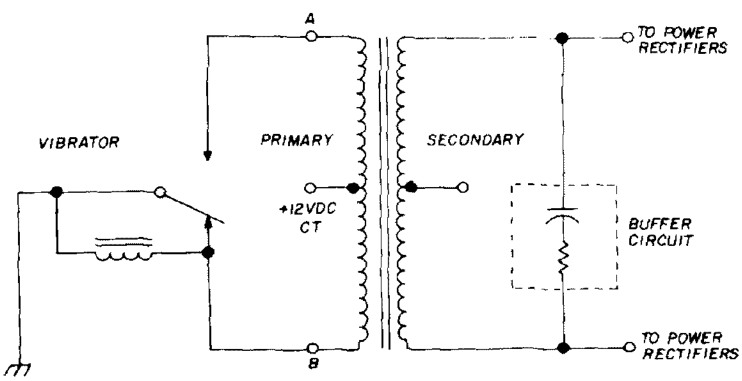


fig. 1. The basic vibrator supply.

Construction is not critical. The tran-

sistors should be heat sinked as they do run warm under transmit condition. Other components should be mounted with care keeping vibration in mind. A short barrier strip mounted on the outside of the box would permit quick connection and testing without connection to the rig. The switch can be tested without connection to the rig by applying plus 12 Vdc to pin four of the toroid and providing a ground return. The switch will oscillate at a low frequency (you'll hear a buzz from the toroid); however, when a load is connected, the operating frequency of the switch is near 1000 Hz. If the load becomes too great, the circuit will stop oscillating and save the transistors from thermal runaway.

Since the switch has only two connections to the rig plus the chassis ground, the hookup time is minimal. The vibrator is removed from its socket. The base of an old vibrator could be used as a quick connector to the vibrator socket. The two pins to be used should be determined from the schematic of your rig. Run two heavy leads from terminals A and B to the pins that correspond to the ends of the primary. A spst switch should now be mounted on the power-supply chassis, accessible from the top side of the chassis. This switch is wired in series with the buffer circuit. Be sure to label this switch as *buffer in* in the *on* position and *buffer out* in the *off* position. The *buffer out* position is used with the solid-state switch as the reactance of the buffer at the new operation frequency (1000 Hz) is much too low and the power it would pass would burn out the resistor in short order. The rig now is ready to operate with the option of returning to vibrator with the flick of the switch to the *buffer in* position and reinstalling a vibrator.

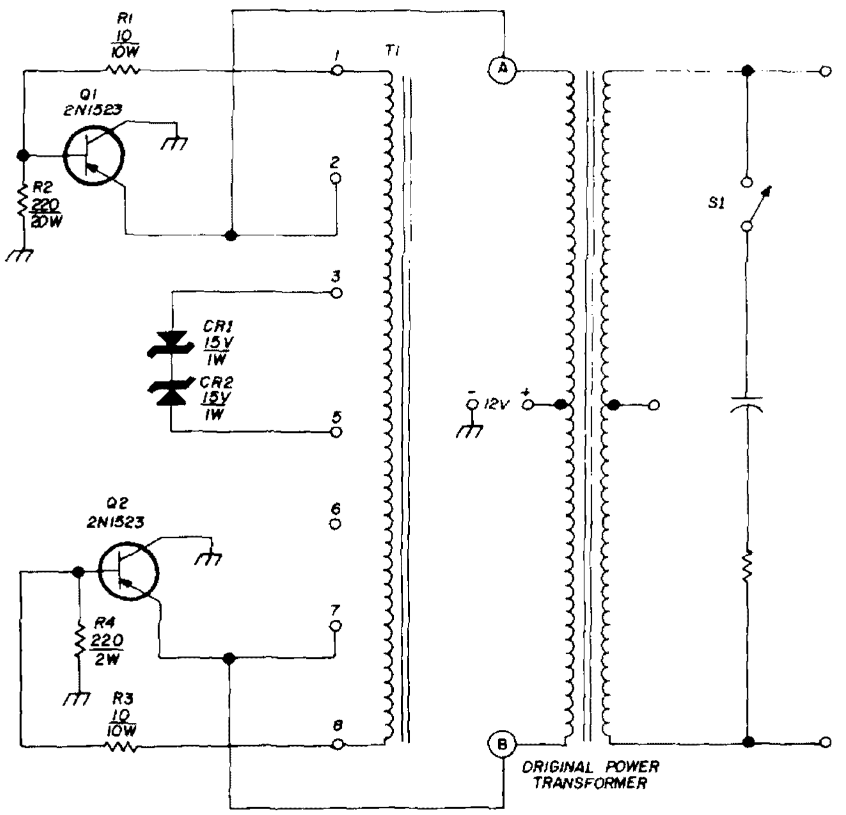


fig. 2. Transistor switch is the heart of the solidstate vibrator replacement. T1 is an Osborne 2709, available from Osborne Transformer Company, 2823 Mitchell Avenue, Detroit, Michigan 48207.

This same basic circuit appeared in the 1970 edition of *The Radio Amateur's Handbook*, page 336, as an integral part of a more complex mobile power supply.

Ray Kashubosky, K8RAY

aligning audio filters

The bandpass characteristics of an audio filter may be best displayed and adjusted through the utilization of an audio sweep system with an oscilloscope. However, such systems are expensive and appear only in the best-equipped laboratories. Consequently, filters are sometimes assembled with complete reliance for their tuning entrusted in the labeled values and tolerances stated on the components.

After constructing the excellent 200-Hz filter described by K7UDL in the September, 1967 issue of *73*, I questioned whether the four 88-mH toroidal coils were accurately labeled. My suspicions were confirmed when five coils were compared and found to exhibit five different values of inductance.

I found a simple method for precisely matching capacitors, inductors and L-C tank tuning which should be helpful in similar applications. I built the two-terminal inductor oscillator shown in fig. 3 and used a high-impedance earphone to monitor the output tone.

Before assembling a filter, all identical L-C tanks are switched in individually, to form the frequency-determining elements of the oscillator. If the audio tones produced by all tanks are not identical in frequency, tank adjustments are necessary. Select matched inductances by switching the coils across one capacitor until you find coils producing the same tone.

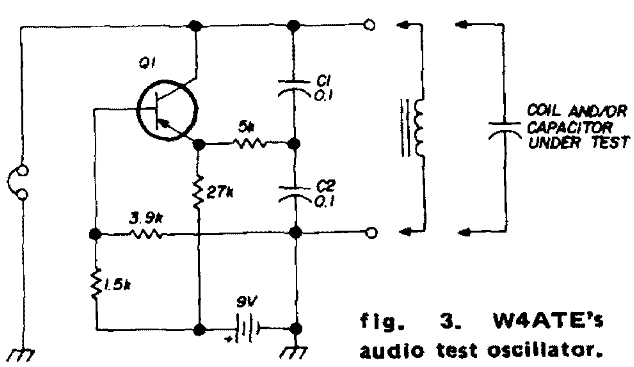


fig. 3. W4ATE's audio test oscillator.

Unpotted coils are convenient, and permit easy adjustment, by removing or adding turns. Match capacitors by switching various capacitor combinations across one toroidal coil until each combination produces exactly the same tone in the earphone. Note that the tone monitored in the earphone does not represent the operating frequency of the assembled filter. The test oscillator serves purely as a comparison device.

In simple one-section filters, tuning may be accomplished by adjustment of either inductance or capacitance. However, in a symmetrical multi-section filter, the Q of equivalent sections should be equal. This requires that both inductance and capacitance in those sections be matched.

With components known to be matched, filters accurately aligned to design frequencies may be assembled with confidence that optimum performance will be enjoyed.

Gene Brizendine, W4ATE

overtone crystals

Do not overtone crystals below 10 Mhz unless you have a frequency meter. The frequency *will not be* three, five, seven or nine times the fundamental. In fact, the frequency can be as much as three percent low. This is shown in fig. 4, a graph of the expected amount of error of low-frequency crystals working on their common overtones. The graph shows average figures taken from tests on many different crystals.

In general, smaller crystals, like the HC-18/U, have bigger errors than larger types, like the HC-6/U or FT-243. However, the older air-mounted types with flat plates (like the 10X) will always be exact multiples.

You can use overtone crystals at their fundamental frequency and at other overtone frequencies. Here is how: Crystal manufacturers can generally make fundamental crystals up to 20 MHz, and they make them for overtones above this. You can easily find out what overtone a crystal works at by dividing the fre-

quency by odd integers (3, 5, 7, 9) until you get the highest frequency *below* 20 MHz. For example, a 75-MHz crystal cannot be a third overtone because 75 MHz divided by three is 25 MHz. This is too

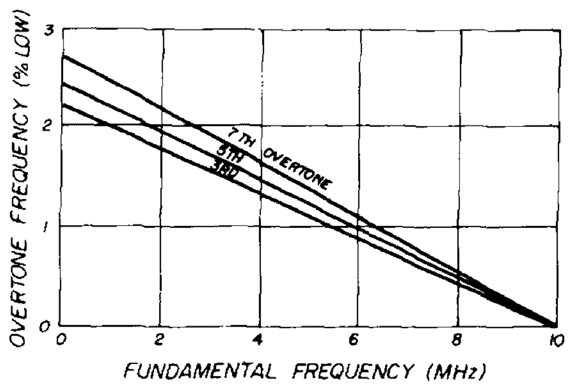


fig. 4. Graph of the projected frequency error when overtoning crystals below 10 MHz.

high, being above 20 MHz. It is a fifth-overtone crystal, as 75 MHz divided by five is 15 MHz. This same crystal can be operated at its third overtone to produce 45 MHz, or on its fundamental to get 15 MHz.

If a particular crystal will overtone at all (some very old ones will not), the Butler circuit will do it. Also use the Butler if you want a high overtone like the ninth. Otherwise the Pierce circuit is fine for good crystals at low overtones.

Martin Mann, G8ABR

key and vox clicks

A casual review of Official Observer logs suggests that there are about four times as many transmitters with key clicks on the *make* of the contacts, than on the *break*. Often, the amount is not large, but it can be heard several kilohertz to each side of the signal causing unnecessary interference. It may not be noticed by a station tuned to receive the signal because the beat note may obscure it.

Recently, a local amateur telephoned about his signal, so we worked the matter out. He uses a Swan 500C with the keying circuit shown in fig. 5 (except for the plug and key). You will see that capacitor C1507 charges when the key is opened, the rate of charge being limited by R1608. However, when the key is closed, the capacitor discharges abruptly, possibly through contact-bounce in the key. This would tend to make the click mushier.

We connected a resistor decade box in series with the key. When it was set at about 10 kilohms, the scope pattern showed a very nice sloping rise which was very similar to the sloping fall at the other end of dots and dashes. The click was entirely gone.

There is *no* modification to the transceiver. The resistor can be mounted on the key or, in many cases, inside of the key plug.

It is probable that this same approach will apply to other exciters, many of which have a filter that is adjustable on one end of dots and dashes, but not at the other end.

Some stations have a *make* click that occurs only after a pause, such as in the first character of a word. Many of these are the result of using a fast vox adjust-

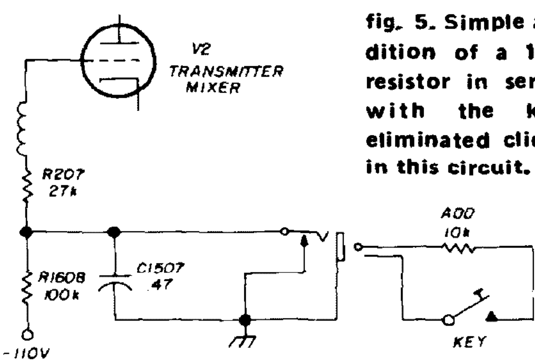
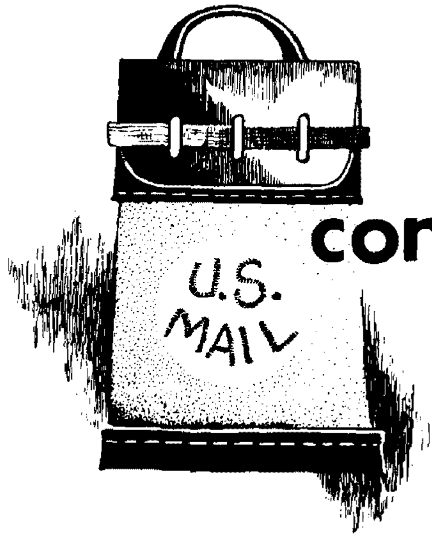


fig. 5. Simple addition of a 10k resistor in series with the key eliminated clicks in this circuit.

ment in an attempt to obtain break-in operation. However, some exciters do not have a key-click filter in the circuit that is operated by the vox. This can cause a click at the *start* of each word sent, but not *within* words unless the keying is *very* slow. Where this occurs, the vox may be slowed down, or a suitable filter can be inserted in the circuit where the vox turns the transmitter on and off. Incidentally, this click also occurs in phone operation.

E. H. Conklin, K6KA



comments

RTTY for the blind

Dear HR:

I recently constructed about a dozen TTL SELCAL units for amateur RTTY. During the course of this activity it has become evident that it would not be too difficult to use similar techniques to build a unit that would make it possible for blind people to read a teletype signal right off the air. If it could be arranged it would open up a new field of interest for our blind friends.

Circuitry can readily be arranged which will recognize any teletype character or letter and, with suitable processing, turn on a light, energize a relay, or as in our case, operate a solenoid. The idea we are presenting is to arrange ten such small solenoids (for use in the case of 5 level teletype) so that 5 of them would be continuously operated by the incoming characters in the LTRS mode of operation while the other five would be operated by the FIGS. The receiving operator could place his left fingers over one of these sets and his right fingers over the other.

It should not take too much practice for the receiving operator to be able to recognize the characters as they came in from the various "highs" and "lows" in each character. Initial practicing could be done at any slow speed such as five or ten words per minute by the simple expedient of having a helper punch occasional letters on a keyboard connected to the "reader." Possibly punched tapes could be played back at any speed the operator chose through the use of a dc speed-controlled motor belted to the standard tape reader motor. With a few month's practice some operators should be able to copy at the full rate of sixty words per minute because each character would be available both to feel and hear instantly. Any necessary combination of characters and blank spaces could be utilized to assist slower operators.

There is a possibility that a whole new area of interest might develop from these suggestions. Blind people could write back and forth to each other on punched tapes. Articles and even whole books could be made available on punched tape. In some cases ordinary magnetic tape could be used to record the familiar audio tones of the RTTY signals and the receiving operator (with a demodulator, modified SELCAL, and our suggested solenoid combination) could "read" the tape. The most useful purpose we are thinking of for the moment, though, is that blind hams with the proper equipment could communicate with any other ham using teletype.

The diagrams will show briefly the techniques we are suggesting. Possibly other ideas from interested parties would help to realize a satisfactory arrangement. While thinking on this subject, it might be kept in mind that people who are deaf could communicate with each other or with blind persons or with anyone else through RTTY circuits. Anyone working on these ideas would be wise to consider the whole concept.

Larry Walrod, VE7BRK
Kelowna, B. C., Canada

diode surge protection

Dear HR:

In the article "Diode Surge Protection" by John Lapham, WA7LUJ, which appeared on page 65 of the March, 1972 issue of *ham radio*, the author is correct when he is talking about high-voltage power supplies using modern diodes – the transformer's secondary dc resistance is usually sufficient to protect the bridge diodes against surge current. However, in low-voltage supplies it is a critical problem which should not be overlooked.

An example is a power supply I was designing for my workbench. I had a 10 V transformer rated at 4 A. It had a secondary dc resistance of 0.33 ohms. With the other components on hand – an epoxy bridge rectifier and a 2000 mF capacitor – I had a 50 A surge current for one duty cycle and I needed 0.6 ohms of surge resistance to limit the surge current. Without the addition of an external resistor, I would be keeping the diode manufacturers in business.

It only takes a few seconds to make the surge calculations which could save lots of headaches and pennies later. I recommend it for any design, but especially in low-voltage designs.

Mr. Lapham stated, "Surge protection is a needless feature in power supplies." This is simply not so in low-voltage designs, and I think this distinction should be made clear.

Joseph P. Bremmer, WB6KXF
Pomona, California

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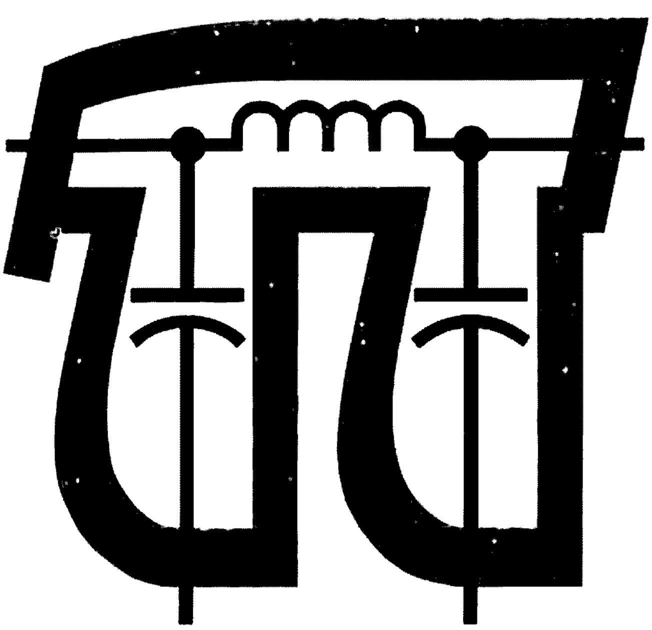
SEPTEMBER 1972

● high-frequency power amplifier PI NETWORK DESIGN



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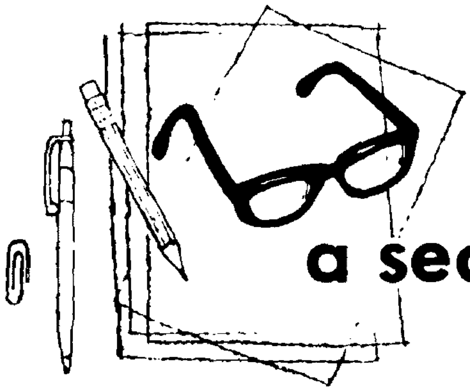

a second look

by jim fisk

A whole new family of ICs which is blooming in the development stage promises to revolutionize communications. These new ICs, called OICs (pronounced *oyks*) for optical integrated circuits, use technology borrowed directly from the semiconductor IC industry.

OICs are closely related to laser communications, and provide lasing, coupling, modulating, filtering and multiplexing for future optical communications systems. Practical application of these new devices is anywhere from two to ten years in the future, depending on who you talk to. However, it is just a matter of time before optical communications systems carry the bulk of industrial traffic, including telephone, digital data and television. The big advantage of the optical system, of course, is the huge message-carrying bandwidth which is available.

Early optical communications systems depend upon short laser links, but in future systems there will give way to fiber-optic *wires* which carry the signal from point to point. The optical fiber, already developed by Corning Glass, will carry a coherent laser signal with relatively small losses. OICs will be used to process the optical signal once it reaches its destination.

Both active and passive types of OICs are being developed at the present time. The most successful passive types use a thin glass film deposited on a glass plate. A masking and etching process similar to that used by IC manufacturers removes the excess glass deposit. These passive structures can be formed into frequency-selective filters and directional couplers as well as straight signal-carrying optical waveguides.

Development work on active OICs is progressing on several different, divergent fronts. Some scientists are fabricating optical waveguides in gallium-arsenide compounds, a familiar semiconductor material. Waveguide effects have also been produced in zinc selenide diffused in cadmium. The materials and processes being developed for OICs remind me of the then-new and strange semiconductor processes which spawned the transistor industry in the late 1950s.

Many problems will have to be solved before optical communications and OICs become a practical reality, including a method of putting laser energy into the device and taking it out. Two methods that show promise include prisms placed very close to the surface and optical gratings etched on the waveguide surface. Prisms are much more efficient than the gratings, but they cost considerably more and are more difficult to build. However, research is progressing in this area, and in all probability, a low-loss grating will eventually be developed.

Although few amateurs have added laser-communications to their repertoire, some amateurs have had good success, including the two-way laser phone contact of WA8EWJ and W4UDS, and W4KAE's nearly 4-mile QSO with a keyed laser beam. With the availability of new optical equipment, including OICs, range will increase, and someday, in the not too distant future, we may even have an amateur band in the visible light range. Anyone for a schedule on 500 Terahertz?

Jim Fisk, W1DTY
editor

high-frequency power amplifier pi network design

A complete discussion of pi and pi-L network design, with computer-derived component values for a wide range of operating conditions

Irvin M. Hoff, W6FFC, 12130 Foothill Lane, Los Altos Hills, California 94022

The design of rf power amplifiers has always fascinated the typical radio amateur, and it remains one of the few fields in which a person of modest technical capability can still actively participate. Although the number of home-built transmitters has steadily diminished as more commercial companies have entered the market, many amateurs still like to design and build their own final amplifier. The information contained in this article should greatly assist those so inclined. Many interesting comparisons will be presented between amplifiers running at different power levels as well as pertinent computer-derived data for the proper selection of component values.

With single sideband and its legal 2-kW PEP maximum input power, certain problems crop up which many amateurs overlook or are unable to handle. This is because the operator wants to run the amplifier at one power level for ssb and another for CW. The problems are compounded when the operator also wants to run RTTY, which is 100% key-down continuous-carrier operation.

There is also a growing tendency to build power amplifiers with higher plate voltages than were common a few years ago. Part of this trend is due to the fact that the newer power tubes provide maximum performance at high plate voltages. Many of the pi-network design charts previously published have not been extended to include these higher operating voltages.

The pi network is so named because of its resemblance to the Greek letter pi as shown in **fig. 1**. The same network in its electrical form with input and output impedances is shown in **fig. 2**. Since most amateurs use 50-ohm *coaxial transmission* line, the output load impedance of the pi network is usually 50 ohms.

When the pi network is used in a power amplifier, the circuit looks like that shown in **fig. 3**. The antenna provides the output load impedance, Z_L, and the power tube provides the input load impedance Z_p. Since the plate load impedance usually falls into the range of 1200 to 5000 ohms, the pi network transforms the high impedance of the vacuum tube into the 50-ohm antenna load. It performs this job quite efficiently, and with predictable results.

harmonic attenuation

Actually, the pi network is a basic form of three-pole low-pass filter. With proper care in design it will attenuate the second harmonic by 35 dB or more.[1] This would be for a loaded Q of 12; if the Q is doubled, attenuation is increased by approximately 6 dB.

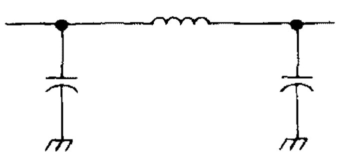

fig. 1. The pi network is so named because of its basic resemblance to the Greek letter π.

The pi-L network shown in **fig. 4** consists of a standard pi network with an additional inductor. Since the pi-L network is a four-pole low-pass filter, second harmonic attenuation is increased to approximately 50 dB. This is particularly important if you want maximum suppression of TVI.

In addition to increased harmonic suppression, the pi-L network offers greater bandwidth for a given variation in operating Q, requires less output capacitance, and is able to operate efficiently with lower Q at very high plate load impedances. These advantages will become more apparent later in this article.

plate load impedance

The dc plate resistance of a vacuum tube, at a given input power level, can be calculated with Ohm's law: $R = E/I$, where E is the dc plate voltage and I is the dc plate current. However, since we are dealing with an ac circuit, this is of little value. What we need to know is the plate load *impedance*. This is given approximately by the following equation which has been derived from the complex functions of a vacuum tube operating in class B.

$$Z_p \approx \frac{E}{1.57\,I} \quad (1)$$

where Z_p is the plate load impedance, I is the indicated plate current and E is the dc plate voltage.

When the vacuum tube is operated in class C, as for CW, the plate load impedance is approximated by

$$Z_p \approx \frac{E}{2 \cdot I} \quad (2)$$

If you are using a linear amplifier that runs with very high idling current, and approaches class A, the following approximation for plate load impedance would be more appropriate.

$$Z_p \approx \frac{E}{1.3\,I} \quad (3)$$

Zero-bias grounded-grid linears are usually thought of as being class B, but there is no hard and fast rule in this regard. A number of articles have been written on this subject, and you are likely to have already formed some opinions of your own.

Consider the case of a class-B rf power amplifier with a 2100-volt plate power supply and indicated plate current of 476 mA (1 kW input). As calculated from **eq. 2**, the plate load impedance is 2800 ohms:

$$Z_p = \frac{2100}{1.57 \cdot 0.476} = 2800 \text{ ohms}$$

Typical plate load impedances for various power levels and different operating

voltages and currents are shown in **table 1**. It can be seen from this data that the plate load impedance rises to very high levels when the plate voltage is increased above 4000 volts. More amateurs than might be expected use 4000 to 6000 volt power supplies, and many of the associated problems have not been adequately discussed in the past.

circuit Q

The letter Q stands for quality factor, and is used to describe, in simple numerical terms, the efficiency and performance of capacitors and inductors. Actually, there are two types of Q – *loaded* Q and *unloaded* Q. The unloaded Q is the inherent quality factor of the component itself; loaded Q is the quality factor of the component when it is used (and loaded down) by the circuit.

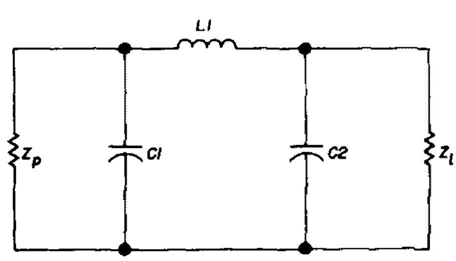

fig. 2. Basic pi network showing the input and output load impedances. The input load impedance in transmitters is the plate load impedance; output load impedance is usually 50 ohms in amateur stations.

The unloaded Q of a component is given by

$$Q_u = Q\ (\text{unloaded}) = \frac{X}{r} \qquad (4)$$

where X is reactance and r is ac resistance. The unloaded Q of a high-quality capacitor might be 1000 or more, and a silver-plated inductor might have an unloaded Q of more than 500.

The loaded Q of a pi network is usually on the order of 10 to 20 for maximum harmonic attenuation, and is given by:

$$Q_o = Q\ (\text{loaded}) = \sqrt{\frac{Z_p \cdot Z_L}{Z_L}}$$

where Z_p is the input impedance to the network, and Z_L is the output impedance.

When designing pi networks a value of

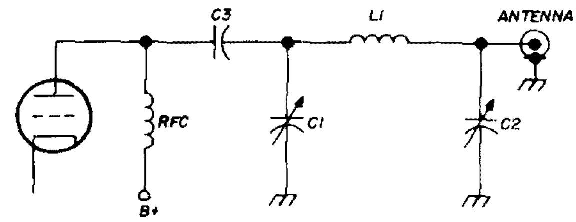

fig. 3. Pi network used in the output of a rf power amplifier is coupled to the power tube through a dc blocking capacitor (C3). C1 is the tuning capacitor, C2 is the loading capacitor, and L1 is the tank inductor.

loaded Q is chosen on the basis of harmonic attenuation, and is used in the design equations to determine the inductance and capacitance values for a given operating frequency.

L networks

A typical step-down L network is shown in fig. 5. This network is used to transform its input impedance to a lower output impedance. The Q of this circuit is entirely dependent upon the ratio of the input and output impedances as given in **eq. 5**.

For example, if the input impedance to an L network is 2500 ohms, and the output impedance is 50 ohms, the loaded Q of the network is 7:

$$Q_o = Q\ (\text{loaded}) = \sqrt{\frac{2500 - 50}{50}} = \sqrt{49} = 7$$

However, a loaded Q of 7 is much too low for good harmonic suppression. To determine the L-network input im-

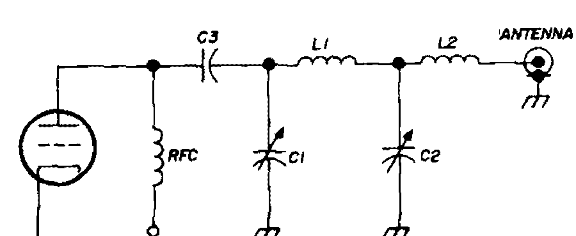

fig. 4. The pi-L network requires an additional inductor, and provides increased second harmonic attenuation.

pedance required to provide a desired value of loaded Q, eq. 5 is rearranged as shown below:

$$Z_p \approx Z_L(Q_o^2 + 1) \qquad (6)$$

For example, with an output load impedance of 50 ohms, and a desired loaded Q of 12 (for good harmonic suppression), the required input impedance is 7250 ohms. This is very restrictive and does not allow the designer sufficient latitude. So, although the L network is extremely efficient (98% typical), a pi network is usually used in transmitter output circuits.

pi network design

You can think of the pi network as being two L networks in tandem as shown in fig. 6. The first L network is a step-down type while the second L network is reversed for impedance step up. As an example, consider the case where the input impedance to the dissected pi network in fig. 6 is 2900 ohms. With a Q of 12, the first L network would step the input impedance down to 20 ohms. This is often called the *virtual* impedance.

$$Z_L = \frac{Z_p}{Q_o^2 + 1} = \frac{2900}{12^2 + 1} = \frac{2900}{145} = 20 \text{ ohms} \qquad (7)$$

The second L network would then be designed to raise this virtual impedance of 20 ohms to 50 ohms to match the antenna. The Q of the second section would be quite low, on the order of 1.5.

As the input impedance is increased with Q held constant, the virtual impedance increases, and when the virtual impedance is equal to the desired output impedance, the pi network reverts to an L network. For example, with a plate load impedance of 7250 ohms and a Q of 12, the virtual impedance is 50 ohms. This is the maximum possible *impedance* transformation for a Q of 12 and an output impedance of 50 ohms.

Normally, about 70% of the maximum possible impedance transformation is used in a practical circuit. For a Q of 12 and an antenna load of 50 ohms, this would represent a plate load resistance of 5075 ohms. If the plate load resistance in

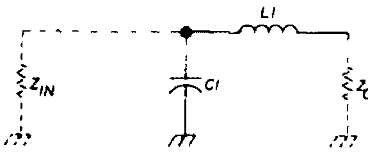

fig. 5. Typical step-down L network is highly efficient but very restrictive as far as acceptable Q is concerned.

a rf power amplifier is higher than 5075 ohms, a Q of more than 12 is required to retain the same level of harmonic suppression. This problem is circumvented by the use of the pi-L network, as discussed below.

pi-L network design

Another L network may be added to

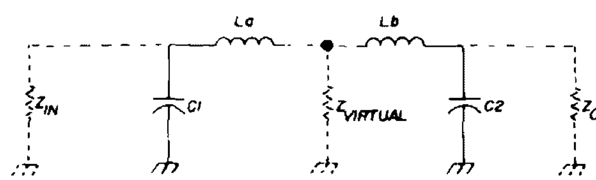

fig. 6. Pi network is basically two L networks in tandem.

the pi network as shown in fig. 7 for additional harmonic attenuation. In actual practice C2 and C3 are combined into *one capacitor* so the circuit used in the transmitter is like that shown in fig. 4.

In the pi-L network, the input pi

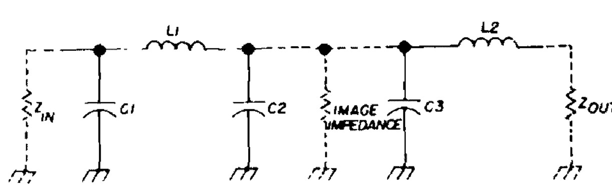

fig. 7. In the pi-L network a second L network is added to the basic pi network. Capacitors C2 and C3 are combined into one capacitor in a practical circuit, as shown in fig. 4.

section transforms the plate load impedance to some lower figure, such as 300 ohms; this is often called the *image* impedance. The final L network transforms the image impedance down to 50 ohms to match the antenna.

From eq. 6 it can be seen that with an image impedance of 300 ohms and a Q of 12, the pi network has a maximum transformation of 43500 ohms. Using 70% of the maximum possible transformation as a practical maximum, as noted before, results in a maximum practical input impedance of 30500 ohms with a Q of 12. This is far in excess of what you will ever need in a power amplifier designed for amateur service.

The image impedance usually falls in the range between 200 and 400 ohms. It is selected for good harmonic attenuation, as well as balance in the T section of the pi-L network and reasonable component values for the capacitors and inductors. If the image impedance is too high, the tuning capacitor (C1) will be too small on 10 and 15 meters, and the two inductors will be very large. Large inductors, of course, increase circulating currents which result in higher losses due to heat.

table 1. Plate load impedances for different input power levels and different operating voltages and currents.

input power (W)	V	mA	plate impedance (ohms)
1000	2000	500	2546
2000	2000	1000	1273
2500	2000	1250	1019
1000	2500	400	3979
2000	2500	800	1989
2500	2500	1000	1592
1000	2800	357	4991
2000	2800	714	2496
2500	2800	893	1996
1000	3300	303	6933
2000	3300	606	3466
2500	3300	758	2773
1000	4000	250	10186
2000	4000	500	5093
2500	4000	625	4074
1000	5000	200	15915
2000	5000	400	7958
2500	5000	500	6366
1000	6000	167	22918
2000	6000	333	11459
2500	6000	417	9167

Q vs frequency

The loaded Q of a pi network (or any tank circuit, for that matter) is equal to its parallel-resonant impedance divided by either the inductive or capacitive reactance of the network. The resonant impedance because the pi network is designed to match the tube operating conditions.

$$Q_o = \frac{Z_p}{X} \qquad (8)$$

The reactance of any inductor is directly proportional to frequency, increasing as the frequency increases. Therefore, from eq. 8 it can be seen that if a particular inductor is used, loaded Q will vary inversely with frequency. As the frequency is lowered, for example, Q is raised a proportionate amount. With this in mind, it is easy to determine the Q for a given network on a different frequency from the following formula:

$$\frac{f1}{f2} \cong \frac{Q2}{Q1} \qquad (9)$$

Where f1 and Q1 are the frequency and Q at one frequency, and f2 and Q2 are at the second, different frequency.

For example, if an 80-meter pi network has a Q of 12 at 4.0 MHz, what is the Q at 3.5 MHz?

$$\frac{4.0}{3.5} = \frac{Q2}{12} \qquad Q2 = 13.7$$

Although the actual loaded Q is somewhat dependent upon the value of plate load impedance used in the circuit, this approximation is accurate within 1%. In the above example, with a plate load impedance of 3000 ohms, Q2 would actually be 13.84.

Since the Q of the network goes up as

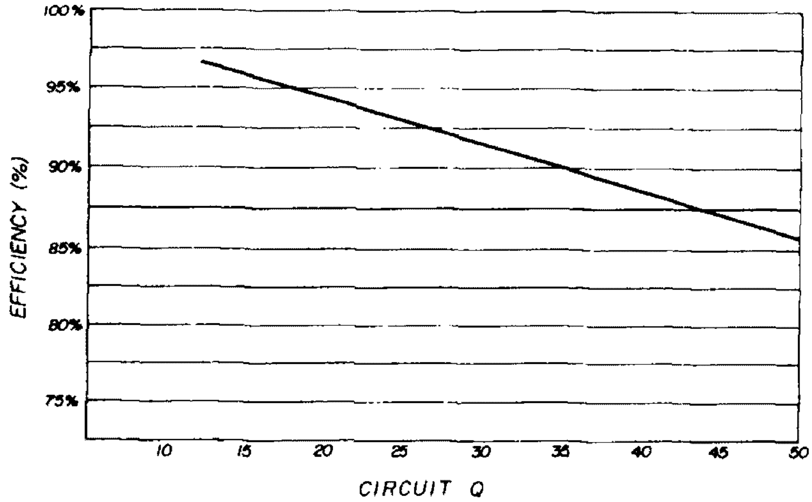

fig. 8. Efficiency of a network is inversely proportional to the loaded Q of the network.

the frequency goes down, it's a good idea to design the pi network for the highest frequency that is to be used.[2] With this approach, when the same inductor is used at lower frequencies within an amateur band, Q increases somewhat, improving harmonic attentuation.

Table 2 shows how Q varies as a pi network is retuned to a different frequency (same inductor). Table 2A shows a pi network designed for 4.0 MHz which is retuned to 3.5 MHz; Q increases from 12 at 4.0 MHz to 13.8 at 3.5 MHz. The values of the tuning and loading capacitors are shown for comparison.

Table 2B shows the case where a pi network is designed for 3.5 MHz with a Q of 12 and retuned to 4.0 MHz (same inductor). The Q drops to 10.4, well below the selected minimum of 12.

network efficiency

As the loaded Q of a network is increased, efficiency goes down because of higher circulating currents and higher losses in the components. Approximate efficiency is given by

$$\text{efficiency} = 100\left(1 - \frac{Q_o}{Q_u}\right) \qquad (10)$$

where Q_o is the loaded circuit Q, and Q_u is the unloaded component Q. The graph in fig. 8 shows that efficiency is a linear function of loaded Q. For minimum loss, the loaded Q should be as low as convenient, while still providing adequate harmonic attenuation. This figure has arbitrarily been chosen as 12.

When the pi network is designed, the minimum Q of 12 can only be obtained at the upper frequency of each amateur band, and then only at the maximum input power level. For other frequencies or lower input powers, the loaded Q is higher than 12.

pi network design

Usually, when you are trying to design a pi network for your transmitter or linear amplifier, you must refer to graphs

table 2. Variations in Q as the resonant frequency of the pi network is changed (same inductor).

	frequency	plate impedance (ohms)	load impedance (ohms)	C1 (pF)	L1 (μH)	C2 (pF)	Q
A. Decreasing	4.0 MHz	2500	50	191	9.18	1097	12.0
frequency	3.5 MHz	2500	50	252	9.18	1536	13.8
B. Increasing	3.5 MHz	2500	50	218	10.49	1254	12.0
frequency	4.0 MHz	2500	50	165	10.49	863	10.4

table 3A. Pi network component values for matching a 50-ohm antenna load. Values have been chosen for a Q of 12 at the top edge of each amateur band. For plate load impedances greater than 5000 ohms, the Q of the network has been adjusted upward to compensate for the maximum transformation ratio, as discussed in the text. R1 is the plate load impedance.

R1 OHMS	F MHZ	C1 PF	L1 UH	C2 PF	R2 OHMS	'Q' QUAL.
200	3.5	3147	0.98	6244	50	13.8
200	7.0	1427	0.54	2827	50	12.6
200	14.0	701	0.27	1387	50	12.3
200	21.0	465	0.18	921	50	12.3
200	29.7	322	0.13	636	50	12.0
300	3.5	2098	1.38	5871	50	13.8
300	7.0	952	0.76	2294	50	12.6
300	14.0	467	0.39	1125	50	12.3
300	21.0	310	0.26	747	50	12.3
300	29.7	214	0.19	516	50	12.0
400	3.5	1573	1.76	4368	50	13.8
400	7.0	714	0.97	1973	50	12.6
400	14.0	350	0.49	968	50	12.3
400	21.0	233	0.33	643	50	12.3
400	29.7	161	0.24	444	50	12.0
500	3.5	1259	2.14	3886	50	13.8
500	7.0	571	1.18	1755	50	12.6
500	14.0	280	0.60	859	50	12.3
500	21.0	186	0.40	571	50	12.3
500	29.7	129	0.29	394	50	12.0
600	3.5	1049	2.51	3528	50	13.8
600	7.0	476	1.38	1590	50	12.6
600	14.0	234	0.70	779	50	12.3
600	21.0	155	0.47	517	50	12.3
600	29.7	107	0.34	357	50	12.0
700	3.5	899	2.88	3248	50	13.8
700	7.0	408	1.58	1462	50	12.6
700	14.0	200	0.81	716	50	12.3
700	21.0	133	0.54	475	50	12.3
700	29.7	92	0.39	328	50	12.0
800	3.5	787	3.24	3021	50	13.8
800	7.0	357	1.78	1358	50	12.6
800	14.0	175	0.91	665	50	12.3
800	21.0	116	0.61	442	50	12.3
800	29.7	80	0.44	304	50	12.0
900	3.5	699	3.60	2832	50	13.8
900	7.0	317	1.98	1271	50	12.6
900	14.0	156	1.01	622	50	12.3
900	21.0	103	0.67	413	50	12.3
900	29.7	71	0.49	285	50	12.0
1000	3.5	629	3.96	2671	50	13.8
1000	7.0	285	2.18	1197	50	12.6
1000	14.0	140	1.11	586	50	12.3
1000	21.0	93	0.74	389	50	12.3
1000	29.7	64	0.54	268	50	12.0
1100	3.5	572	4.32	2532	50	13.8
1100	7.0	260	2.37	1133	50	12.6
1100	14.0	127	1.21	555	50	12.3
1100	21.0	85	0.81	368	50	12.3
1100	29.7	58	0.58	255	50	12.0
1200	3.5	524	4.67	2410	50	13.8
1200	7.0	238	2.57	1077	50	12.6
1200	14.0	117	1.31	527	50	12.3
1200	21.0	78	0.87	350	50	12.3
1200	29.7	54	0.63	241	50	12.0
1300	3.5	484	5.03	2302	50	13.8
1300	7.0	220	2.76	1027	50	12.6
1300	14.0	108	1.40	502	50	12.3
1300	21.0	72	0.94	333	50	12.3
1300	29.7	49	0.68	229	50	12.0
1400	3.5	450	5.38	2205	50	13.8
1400	7.0	204	2.95	982	50	12.6
1400	14.0	100	1.50	480	50	12.3
1400	21.0	66	1.01	319	50	12.3
1400	29.7	46	0.73	219	50	12.0
1500	3.5	420	5.73	2117	50	13.8
1500	7.0	190	3.14	942	50	12.6
1500	14.0	93	1.60	460	50	12.3
1500	21.0	62	1.07	305	50	12.3
1500	29.7	43	0.77	210	50	12.0
2000	3.5	315	7.46	1776	50	13.8
2000	7.0	143	4.09	783	50	12.6
2000	14.0	70	2.08	382	50	12.3
2000	21.0	47	1.39	253	50	12.3
2000	29.7	32	1.01	174	50	12.0
2500	3.5	252	9.17	1536	50	13.8
2500	7.0	114	5.03	670	50	12.6
2500	14.0	56	2.56	326	50	12.3
2500	21.0	37	1.71	216	50	12.3
2500	29.7	26	1.24	148	50	12.0
3000	3.5	210	10.86	1352	50	13.8
3000	7.0	95	5.95	583	50	12.6
3000	14.0	47	3.03	283	50	12.3
3000	21.0	31	2.02	187	50	12.3
3000	29.7	21	1.46	120	50	12.0
3500	3.5	180	12.53	1203	50	13.8
3500	7.0	82	6.86	512	50	12.6
3500	14.0	40	3.49	247	50	12.3
3500	21.0	27	2.34	164	50	12.3
3500	29.7	18	1.69	111	50	12.0
4000	3.5	157	14.19	1079	50	13.8
4000	7.0	71	7.77	451	50	12.6
4000	14.0	35	3.95	217	50	12.3
4000	21.0	23	2.64	144	50	12.3
4000	29.7	16	1.91	97	50	12.0
4500	3.5	140	15.84	971	50	13.8
4500	7.0	63	8.66	397	50	12.6
4500	14.0	31	4.40	190	50	12.3
4500	21.0	21	2.95	126	50	12.3
4500	29.7	14	2.13	84	50	12.0
5000	3.5	126	17.48	875	50	13.8
5000	7.0	57	9.55	348	50	12.6
5000	14.0	28	4.85	165	50	12.3
5000	21.0	19	3.25	109	50	12.3
5000	29.7	13	2.34	72	50	12.0
5500	3.5	119	18.41	861	50	14.4
5500	7.0	54	10.05	341	50	13.1
5500	14.0	27	5.11	162	50	12.8
5500	21.0	18	3.42	107	50	12.8
5500	29.7	12	2.47	70	50	12.5
6000	3.5	114	19.18	862	50	15.1
6000	7.0	52	10.48	341	50	13.7
6000	14.0	25	5.33	162	50	13.4
6000	21.0	17	3.56	107	50	13.4
6000	29.7	12	2.57	70	50	13.1
6500	3.5	110	19.92	862	50	15.7
6500	7.0	50	10.89	341	50	14.2
6500	14.0	24	5.54	162	50	14.0
6500	21.0	16	3.70	107	50	13.9
6500	29.7	11	2.67	70	50	13.6
7000	3.5	106	20.63	862	50	16.3
7000	7.0	48	11.28	341	50	14.8
7000	14.0	24	5.74	162	50	14.5
7000	21.0	16	3.84	107	50	14.4
7000	29.7	11	2.77	70	50	14.1
7500	3.5	102	21.31	862	50	16.8
7500	7.0	46	11.66	341	50	15.3
7500	14.0	23	5.93	162	50	15.0
7500	21.0	15	3.97	107	50	14.9
7500	29.7	10	2.86	70	50	14.6
8000	3.5	99	21.98	862	50	17.4
8000	7.0	45	12.02	341	50	15.8
8000	14.0	22	6.11	162	50	15.5
8000	21.0	15	4.09	107	50	15.4
8000	29.7	10	2.95	70	50	15.1
8500	3.5	96	22.62	862	50	17.9
8500	7.0	44	12.38	341	50	16.3
8500	14.0	21	6.29	162	50	16.0
8500	21.0	14	4.21	107	50	15.9
8500	29.7	10	3.04	70	50	15.6
9000	3.5	93	23.24	862	50	18.5
9000	7.0	42	12.72	341	50	16.7
9000	14.0	21	6.47	162	50	16.4
9000	21.0	14	4.33	107	50	16.4
9000	29.7	10	3.12	70	50	16.0
9500	3.5	91	23.85	862	50	19.0
9500	7.0	41	13.06	341	50	17.2
9500	14.0	20	6.64	162	50	16.9
9500	21.0	13	4.44	107	50	16.8
9500	29.7	9	3.21	70	50	16.4
10000	3.5	88	24.44	862	50	19.5
10000	7.0	40	13.38	341	50	17.7
10000	14.0	20	6.81	162	50	17.3
10000	21.0	13	4.55	107	50	17.3
10000	29.7	9	3.29	70	50	16.9
10500	3.5	86	25.02	862	50	19.9
10500	7.0	39	13.70	341	50	18.1
10500	14.0	19	6.97	162	50	17.8
10500	21.0	13	4.66	107	50	17.7
10500	29.7	9	3.37	70	50	17.3
11000	3.5	84	25.58	862	50	20.4
11000	7.0	38	14.01	341	50	18.5
11000	14.0	19	7.13	162	50	18.2
11000	21.0	12	4.77	107	50	18.1
11000	29.7	9	3.44	70	50	17.7
11500	3.5	83	26.13	862	50	20.9
11500	7.0	37	14.32	341	50	18.9
11500	14.0	18	7.28	162	50	18.6
11500	21.0	12	4.87	107	50	18.5
11500	29.7	8	3.52	70	50	18.1
12000	3.5	81	26.67	862	50	21.3
12000	7.0	37	14.61	341	50	19.3
12000	14.0	18	7.43	162	50	19.0
12000	21.0	12	4.97	107	50	18.9
12000	29.7	8	3.59	70	50	18.5

shown in reference books such as the ARRL "Radio Amateur's Handbook."[3]

These graphs are often somewhat confusing because you must first determine the

table 3B. Pi-network component values for use within the 160-meter amateur band. Values were determined as in table 3A.

R1 OHMS	F MHZ	C1 PF	L1 UH	C2 PF	R2 OHMS	'Q' QUAL.
1000	1.8	1187	7.94	5019	50	13.4
1000	1.9	1062	7.95	4459	50	12.7
1000	2.0	955	7.96	3979	50	12.0
1250	1.8	949	9.71	4419	50	13.4
1250	1.9	849	9.72	3917	50	12.7
1250	2.0	764	9.74	3487	50	12.0
1500	1.8	791	11.47	3969	50	13.4
1500	1.9	708	11.48	3510	50	12.7
1500	2.0	637	11.49	3116	50	12.0
1750	1.8	678	13.21	3613	50	13.4
1750	1.9	607	13.22	3197	50	12.7
1750	2.0	546	13.23	2822	50	12.0
2000	1.8	593	14.94	3322	50	13.4
2000	1.9	531	14.94	2922	50	12.7
2000	2.0	477	14.95	2579	50	12.0
2250	1.8	527	16.65	3076	50	13.4
2250	1.9	472	16.65	2697	50	12.7
2250	2.0	424	16.66	2373	50	12.0
2500	1.8	475	18.36	2864	50	13.4
2500	1.9	425	18.36	2504	50	12.7
2500	2.0	382	18.36	2194	50	12.0
2750	1.8	432	20.05	2678	50	13.4
2750	1.9	386	20.05	2333	50	12.7
2750	2.0	347	20.04	2036	50	12.0
3000	1.8	396	21.74	2513	50	13.4
3000	1.9	354	21.73	2181	50	12.7
3000	2.0	318	21.72	1894	50	12.0
3250	1.8	365	23.41	2364	50	13.4
3250	1.9	327	23.40	2043	50	12.7
3250	2.0	294	23.38	1766	50	12.0
3500	1.8	339	25.08	2228	50	13.4
3500	1.9	303	25.06	1917	50	12.7
3500	2.0	273	25.04	1647	50	12.0
3750	1.8	316	26.75	2104	50	13.4
3750	1.9	283	26.72	1801	50	12.7
3750	2.0	255	26.68	1538	50	12.0
4000	1.8	297	28.40	1988	50	13.4
4000	1.9	265	28.37	1693	50	12.7
4000	2.0	239	28.32	1435	50	12.0
4250	1.8	279	30.05	1881	50	13.4
4250	1.9	250	30.00	1591	50	12.7
4250	2.0	225	29.95	1337	50	12.0
4500	1.8	264	31.69	1779	50	13.4
4500	1.9	236	31.64	1495	50	12.7
4500	2.0	212	31.57	1244	50	12.0
4750	1.8	250	33.33	1684	50	13.4
4750	1.9	224	33.26	1403	50	12.7
4750	2.0	201	33.17	1155	50	12.0
5000	1.8	237	34.96	1593	50	13.4
5000	1.9	212	34.87	1315	50	12.7
5000	2.0	191	34.77	1058	50	12.0
5250	1.8	230	36.01	1566	50	13.7
5250	1.9	206	35.93	1290	50	12.9
5250	2.0	185	35.82	1042	50	12.2
5500	1.8	225	36.81	1566	50	14.0
5500	1.9	201	36.73	1290	50	13.2
5500	2.0	181	36.63	1042	50	12.5
5750	1.8	220	37.59	1567	50	14.3
5750	1.9	197	37.52	1290	50	13.5
5750	2.0	177	37.41	1042	50	12.8
6000	1.8	215	38.36	1567	50	14.6
6000	1.9	193	38.28	1290	50	13.8
6000	2.0	173	38.18	1042	50	13.1
6250	1.8	211	39.11	1567	50	14.9
6250	1.9	189	39.03	1290	50	14.1
6250	2.0	170	38.94	1042	50	13.3
6500	1.8	207	39.84	1567	50	15.2
6500	1.9	185	39.77	1290	50	14.4
6500	2.0	166	39.68	1042	50	13.6
6750	1.8	203	40.56	1567	50	15.5
6750	1.9	182	40.49	1290	50	14.6
6750	2.0	163	40.40	1042	50	13.9
7000	1.8	199	41.27	1567	50	15.8
7000	1.9	178	41.20	1290	50	14.9
7000	2.0	160	41.11	1042	50	14.1
7250	1.8	196	41.96	1567	50	16.1
7250	1.9	175	41.90	1290	50	15.2
7250	2.0	158	41.81	1042	50	14.4
7500	1.8	193	42.64	1567	50	16.3
7500	1.9	172	42.58	1290	50	15.4
7500	2.0	155	42.50	1042	50	14.6
7750	1.8	189	43.31	1567	50	16.6
7750	1.9	169	43.25	1290	50	15.7
7750	2.0	152	43.17	1042	50	14.8
8000	1.8	186	43.97	1567	50	16.9
8000	1.9	167	43.92	1290	50	15.9
8000	2.0	150	43.84	1042	50	15.1
8250	1.8	184	44.62	1567	50	17.1
8250	1.9	164	44.57	1290	50	16.2
8250	2.0	148	44.49	1042	50	15.3
8500	1.8	181	45.26	1567	50	17.4
8500	1.9	162	45.21	1290	50	16.4
8500	2.0	146	45.14	1042	50	15.6
8750	1.8	178	45.89	1567	50	17.6
8750	1.9	160	45.84	1290	50	16.7
8750	2.0	144	45.77	1042	50	15.8
9000	1.8	176	46.51	1567	50	17.9
9000	1.9	157	46.47	1290	50	16.9
9000	2.0	142	46.40	1042	50	16.0
9250	1.8	173	47.13	1567	50	18.1
9250	1.9	155	47.08	1290	50	17.1
9250	2.0	140	47.02	1042	50	16.2
9500	1.8	171	47.73	1567	50	18.4
9500	1.9	153	47.69	1290	50	17.4
9500	2.0	138	47.63	1042	50	16.4
9750	1.8	169	48.33	1567	50	18.6
9750	1.9	151	48.29	1290	50	17.6
9750	2.0	136	48.23	1042	50	16.7
10000	1.8	167	48.92	1568	50	18.9
10000	1.9	149	48.88	1290	50	17.8
10000	2.0	134	48.82	1042	50	16.9
10250	1.8	165	49.50	1568	50	19.1
10250	1.9	147	49.46	1290	50	18.0
10250	2.0	133	49.41	1042	50	17.1
10500	1.8	163	50.07	1568	50	19.3
10500	1.9	146	50.04	1290	50	18.3
10500	2.0	131	49.98	1042	50	17.3
10750	1.8	161	50.64	1568	50	19.6
10750	1.9	144	50.61	1290	50	18.5
10750	2.0	130	50.56	1042	50	17.5
11000	1.8	159	51.20	1568	50	19.8
11000	1.9	142	51.17	1290	50	18.7
11000	2.0	128	51.12	1042	50	17.7
11250	1.8	157	51.76	1568	50	20.0
11250	1.9	141	51.73	1290	50	18.9
11250	2.0	127	51.68	1042	50	17.9
11500	1.8	156	52.30	1568	50	20.2
11500	1.9	139	52.28	1290	50	19.1
11500	2.0	125	52.23	1042	50	18.1
11750	1.8	154	52.85	1568	50	20.5
11750	1.9	138	52.82	1290	50	19.3
11750	2.0	124	52.78	1042	50	18.3
12000	1.8	152	53.38	1568	50	20.7
12000	1.9	136	53.36	1290	50	19.5
12000	2.0	123	53.32	1042	50	18.5

plate load impedance, select a value of Q, and then find the reactance of each of the components. Then you must locate yet another graph to convert these reactance values into actual values of inductance and capacitance.

Few of these charts and graphs are extended above plate load impedances of 5000 ohms, and most give vague reference to the fact that if the plate load impedance *is* greater than about 5000 ohms, the Q should be increased.

The charts in table 3 and table 4 are computer derived and offer all the required information to build a practical pi network. The pi networks in table 3 for plate load impedances above 5000 ohms have increasing Q so that the transformation ratio never exceeds the 70% maximum. *In addition, the inductors chosen* for each of the designs are calculated for the highest frequency in the band.

table 4A. Pi-L network component values for matching a 50-ohm antenna load. Values have been chosen for a Q of 12 at the top edge of each amateur band. The image impedance (R3) has been chosen to provide a balanced transformation in the T section of the pi-L network. R1 is the plate load impedance.

R1 OHMS	F MHZ	C1 PF.	L1 UH.	C2 PF.	L2 UH.	R3 OHMS	R2 OHMS	'Q' QUAL.
500	3.5	1219	2.86	1749	4.45	241	50	13.4
500	7.0	565	1.59	753	2.44	280	50	12.4
500	14.0	277	0.81	365	1.24	288	50	12.2
500	21.0	185	0.54	243	0.83	288	50	12.2
500	29.7	129	0.39	166	0.60	300	50	12.0
600	3.5	1015	3.30	1595	4.45	241	50	13.4
600	7.0	471	1.83	686	2.44	280	50	12.4
600	14.0	231	0.94	332	1.24	288	50	12.2
600	21.0	154	0.63	222	0.83	288	50	12.2
600	29.7	107	0.45	151	0.60	300	50	12.0
700	3.5	870	3.74	1475	4.45	241	50	13.4
700	7.0	403	2.07	635	2.44	280	50	12.4
700	14.0	198	1.06	307	1.24	288	50	12.2
700	21.0	132	0.71	205	0.83	288	50	12.2
700	29.7	92	0.51	140	0.60	300	50	12.0
800	3.5	762	4.17	1378	4.45	241	50	13.4
800	7.0	353	2.31	593	2.44	280	50	12.4
800	14.0	173	1.18	287	1.24	288	50	12.2
800	21.0	116	0.79	191	0.83	288	50	12.2
800	29.7	80	0.57	130	0.60	300	50	12.0
900	3.5	677	4.60	1297	4.45	241	50	13.4
900	7.0	314	2.54	558	2.44	280	50	12.4
900	14.0	154	1.30	270	1.24	288	50	12.2
900	21.0	103	0.87	180	0.83	288	50	12.2
900	29.7	71	0.63	123	0.60	300	50	12.0
1000	3.5	609	5.02	1229	4.45	241	50	13.4
1000	7.0	282	2.77	529	2.44	280	50	12.4
1000	14.0	139	1.42	256	1.24	288	50	12.2
1000	21.0	92	0.95	171	0.83	288	50	12.2
1000	29.7	64	0.68	116	0.60	300	50	12.0
1100	3.5	554	5.43	1171	4.45	241	50	13.4
1100	7.0	257	3.00	504	2.44	280	50	12.4
1100	14.0	126	1.53	244	1.24	288	50	12.2
1100	21.0	84	1.02	163	0.83	288	50	12.2
1100	29.7	58	0.74	111	0.60	300	50	12.0
1200	3.5	508	5.85	1120	4.45	241	50	13.4
1200	7.0	235	3.23	482	2.44	280	50	12.4
1200	14.0	116	1.65	233	1.24	288	50	12.2
1200	21.0	77	1.10	156	0.83	288	50	12.2
1200	29.7	54	0.80	106	0.60	300	50	12.0
1300	3.5	469	6.25	1075	4.45	241	50	13.4
1300	7.0	217	3.45	462	2.44	280	50	12.4
1300	14.0	107	1.76	224	1.24	288	50	12.2
1300	21.0	71	1.18	149	0.83	288	50	12.2
1300	29.7	49	0.85	102	0.60	300	50	12.0
1400	3.5	435	6.66	1034	4.45	241	50	13.4
1400	7.0	202	3.67	445	2.44	280	50	12.4
1400	14.0	99	1.88	216	1.24	288	50	12.2
1400	21.0	66	1.25	144	0.83	288	50	12.2
1400	29.7	46	0.90	98	0.60	300	50	12.0
1500	3.5	406	7.06	998	4.45	241	50	13.4
1500	7.0	188	3.89	430	2.44	280	50	12.4
1500	14.0	92	1.99	208	1.24	288	50	12.2
1500	21.0	62	1.33	139	0.83	288	50	12.2
1500	29.7	43	0.96	95	0.60	300	50	12.0
2000	3.5	305	9.05	859	4.45	241	50	13.4
2000	7.0	141	4.97	370	2.44	280	50	12.4
2000	14.0	69	2.54	179	1.24	288	50	12.2
2000	21.0	46	1.69	119	0.83	288	50	12.2
2000	29.7	32	1.22	81	0.60	300	50	12.0
2500	3.5	244	10.99	764	4.45	241	50	13.4
2500	7.0	113	6.03	329	2.44	280	50	12.4
2500	14.0	55	3.08	159	1.24	288	50	12.2
2500	21.0	37	2.05	106	0.83	288	50	12.2
2500	29.7	26	1.48	72	0.60	300	50	12.0
3000	3.5	203	12.90	693	4.45	241	50	13.4
3000	7.0	94	7.07	298	2.44	280	50	12.4
3000	14.0	46	3.61	144	1.24	288	50	12.2
3000	21.0	31	2.41	96	0.83	288	50	12.2
3000	29.7	21	1.74	66	0.60	300	50	12.0
3500	3.5	174	14.79	638	4.45	241	50	13.4
3500	7.0	81	0.10	274	2.44	280	50	12.4
3500	14.0	40	4.13	133	1.24	288	50	12.2
3500	21.0	26	2.76	89	0.83	288	50	12.2
3500	29.7	18	1.99	60	0.60	300	50	12.0
4000	3.5	152	16.67	593	4.45	241	50	13.4
4000	7.0	71	9.12	255	2.44	280	50	12.4
4000	14.0	35	4.65	123	1.24	288	50	12.2
4000	21.0	23	3.10	82	0.83	288	50	12.2
4000	29.7	16	2.24	56	0.60	300	50	12.0
4500	3.5	135	18.53	555	4.45	241	50	13.4
4500	7.0	63	10.12	239	2.44	280	50	12.4
4500	14.0	31	5.17	116	1.24	288	50	12.2
4500	21.0	21	3.44	77	0.83	288	50	12.2
4500	29.7	14	2.49	53	0.60	300	50	12.0
5000	3.5	122	20.37	523	4.45	241	50	13.4
5000	7.0	56	11.13	225	2.44	280	50	12.4
5000	14.0	28	5.68	109	1.24	288	50	12.2
5000	21.0	18	3.78	73	0.83	288	50	12.2
5000	29.7	13	2.73	50	0.60	300	50	12.0
5500	3.5	111	22.21	496	4.45	241	50	13.4
5500	7.0	51	12.12	213	2.44	280	50	12.4
5500	14.0	25	6.18	103	1.24	288	50	12.2
5500	21.0	17	4.12	69	0.83	288	50	12.2
5500	29.7	12	2.97	47	0.60	300	50	12.0
6000	3.5	102	24.03	472	4.45	241	50	13.4
6000	7.0	47	13.11	203	2.44	280	50	12.4
6000	14.0	23	6.69	98	1.24	288	50	12.2
6000	21.0	15	4.46	65	0.83	288	50	12.2
6000	29.7	11	3.22	45	0.60	300	50	12.0
6500	3.5	94	25.85	450	4.45	241	50	13.4
6500	7.0	43	14.09	194	2.44	280	50	12.4
6500	14.0	21	7.19	94	1.24	288	50	12.2
6500	21.0	14	4.79	62	0.83	288	50	12.2
6500	29.7	10	3.46	43	0.60	300	50	12.0
7000	3.5	87	27.65	431	4.45	241	50	13.4
7000	7.0	40	15.07	185	2.44	280	50	12.4
7000	14.0	20	7.69	90	1.24	288	50	12.2
7000	21.0	13	5.12	60	0.83	288	50	12.2
7000	29.7	9	3.70	41	0.60	300	50	12.0
7500	3.5	81	29.45	413	4.45	241	50	13.4
7500	7.0	38	16.05	178	2.44	280	50	12.4
7500	14.0	18	8.18	86	1.24	288	50	12.2
7500	21.0	12	5.45	57	0.83	288	50	12.2
7500	29.7	9	3.93	39	0.60	300	50	12.0
8000	3.5	76	31.25	397	4.45	241	50	13.4
8000	7.0	35	17.02	171	2.44	280	50	12.4
8000	14.0	17	8.68	83	1.24	288	50	12.2
8000	21.0	12	5.78	55	0.83	288	50	12.2
8000	29.7	8	4.17	38	0.60	300	50	12.0
8500	3.5	72	33.03	383	4.45	241	50	13.4
8500	7.0	33	17.99	165	2.44	280	50	12.4
8500	14.0	16	9.17	80	1.24	288	50	12.2
8500	21.0	11	6.11	53	0.83	288	50	12.2
8500	29.7	8	4.41	36	0.60	300	50	12.0
9000	3.5	68	34.82	369	4.45	241	50	13.4
9000	7.0	31	18.95	159	2.44	280	50	12.4
9000	14.0	15	9.66	77	1.24	288	50	12.2
9000	21.0	10	6.44	51	0.83	288	50	12.2
9000	29.7	7	4.64	35	0.60	300	50	12.0
9500	3.5	64	36.59	357	4.45	241	50	13.4
9500	7.0	30	19.91	154	2.44	280	50	12.4
9500	14.0	15	10.15	74	1.24	288	50	12.2
9500	21.0	10	6.77	50	0.83	288	50	12.2
9500	29.7	7	4.88	34	0.60	300	50	12.0
10000	3.5	61	38.36	345	4.45	241	50	13.4
10000	7.0	28	20.87	149	2.44	280	50	12.4
10000	14.0	14	10.64	72	1.24	288	50	12.2
10000	21.0	9	7.09	48	0.83	288	50	12.2
10000	29.7	6	5.11	33	0.60	300	50	12.0
10500	3.5	58	40.13	334	4.45	241	50	13.4
10500	7.0	27	21.82	144	2.44	280	50	12.4
10500	14.0	13	11.12	70	1.24	288	50	12.2
10500	21.0	9	7.42	46	0.83	288	50	12.2
10500	29.7	6	5.34	32	0.60	300	50	12.0
11000	3.5	55	41.89	324	4.45	241	50	13.4
11000	7.0	26	22.78	140	2.44	280	50	12.4
11000	14.0	13	11.61	67	1.24	288	50	12.2
11000	21.0	8	7.74	45	0.83	288	50	12.2
11000	29.7	6	5.58	31	0.60	300	50	12.0
11500	3.5	53	43.65	315	4.45	241	50	13.4
11500	7.0	25	23.73	135	2.44	280	50	12.4
11500	14.0	12	12.09	65	1.24	288	50	12.2
11500	21.0	8	8.06	44	0.83	288	50	12.2
11500	29.7	6	5.81	30	0.60	300	50	12.0
12000	3.5	51	45.40	306	4.45	241	50	13.4
12000	7.0	24	24.67	131	2.44	280	50	12.4
12000	14.0	12	12.57	64	1.24	288	50	12.2
12000	21.0	8	8.38	42	0.83	288	50	12.2
12000	29.7	5	6.04	29	0.60	300	50	12.0

The Q of the network at the highest frequency is 12 except when the plate load impedance is greater than 5075 ohms. The chart shows the capacitance values required to resonate the network to the lowest frequency in the band (maximum capacitance), as well as the operating Q at that frequency. In table 4, the image impedance (R3) at the lower frequency is also given.

table 4B. Pi-L network compenent values for the 160-meter amateur band. Values were determined as in table 4A.

A suitable coil for the L-network inductor consists of two inches of Air-Dux 1606T (6 turns-per-inch, no. 14 wire, 2" diameter). This inductor should be placed at right angles to the main pi inductor to avoid mutual inductance. In the following chart, 7.125 would be slightly over 7 full turns, 2.875 is slightly less than 3 *full* turns.

band	turns	approximate inductance
80	11.000	4.43 μH
40	7.125	2.43 μH
20	4.5000	1.23 μH
15	3.500	0.83 μH
10	2.875	0.60 μH

R1 OHMS	F MHZ	C1 PF.	L1 UH.	C2 PF.	L2 UH.	R3 OHMS	R2 OHMS	'Q' QUAL.
1000	1.8	1155	10.08	2279	8.90	252	50	13.1
1000	1.9	1047	10.12	1978	8.90	275	50	12.5
1000	2.0	955	10.16	1729	8.90	300	50	12.0
1250	1.8	924	12.14	2033	8.90	252	50	13.1
1250	1.9	838	12.18	1764	8.90	275	50	12.5
1250	2.0	764	12.22	1542	8.90	300	50	12.0
1500	1.8	770	14.17	1850	8.90	252	50	13.1
1500	1.9	698	14.20	1606	8.90	275	50	12.5
1500	2.0	637	14.23	1404	8.90	300	50	12.0
1750	1.8	660	16.17	1708	8.90	252	50	13.1
1750	1.9	599	16.19	1482	8.90	275	50	12.5
1750	2.0	546	16.22	1296	8.90	300	50	12.0
2000	1.8	578	18.14	1593	8.90	252	50	13.1
2000	1.9	524	18.15	1382	8.90	275	50	12.5
2000	2.0	477	18.17	1208	8.90	300	50	12.0
2250	1.8	514	20.09	1497	8.90	252	50	13.1
2250	1.9	466	20.10	1299	8.90	275	50	12.5
2250	2.0	424	20.11	1136	8.90	300	50	12.0
2500	1.8	462	22.02	1416	8.90	252	50	13.1
2500	1.9	419	22.02	1229	8.90	275	50	12.5
2500	2.0	382	22.02	1074	8.90	300	50	12.0
2750	1.8	420	23.94	1346	8.90	252	50	13.1
2750	1.9	381	23.93	1168	8.90	275	50	12.5
2750	2.0	347	23.92	1021	8.90	300	50	12.0
3000	1.8	385	25.85	1285	8.90	252	50	13.1
3000	1.9	349	25.83	1115	8.90	275	50	12.5
3000	2.0	318	25.81	975	8.90	300	50	12.0
3250	1.8	356	27.74	1230	8.90	252	50	13.1
3250	1.9	322	27.71	1068	8.90	275	50	12.5
3250	2.0	294	27.68	933	8.90	300	50	12.0
3500	1.8	330	29.62	1182	8.90	252	50	13.1
3500	1.9	299	29.58	1026	8.90	275	50	12.5
3500	2.0	273	29.54	897	8.90	300	50	12.0
3750	1.8	308	31.50	1138	8.90	252	50	13.1
3750	1.9	279	31.45	988	8.90	275	50	12.5
3750	2.0	255	31.40	864	8.90	300	50	12.0
4000	1.8	289	33.37	1099	8.90	252	50	13.1
4000	1.9	262	33.30	954	8.90	275	50	12.5
4000	2.0	239	33.24	834	8.90	300	50	12.0
4250	1.8	272	35.23	1062	8.90	252	50	13.1
4250	1.9	246	35.15	922	8.90	275	50	12.5
4250	2.0	225	35.08	806	8.90	300	50	12.0
4500	1.8	257	37.08	1029	8.90	252	50	13.1
4500	1.9	233	36.99	893	8.90	275	50	12.5
4500	2.0	212	36.91	781	8.90	300	50	12.0
4750	1.8	243	38.92	999	8.90	252	50	13.1
4750	1.9	221	38.83	867	8.90	275	50	12.5
4750	2.0	201	38.73	758	8.90	300	50	12.0
5000	1.8	231	40.76	970	8.90	252	50	13.1
5000	1.9	209	40.65	842	8.90	275	50	12.5
5000	2.0	191	40.54	736	8.90	300	50	12.0
5250	1.8	220	42.60	944	8.90	252	50	13.1
5250	1.9	200	42.47	819	8.90	275	50	12.5
5250	2.0	182	42.35	716	8.90	300	50	12.0
5500	1.8	210	44.43	919	8.90	252	50	13.1
5500	1.9	190	44.29	798	8.90	275	50	12.5
5500	2.0	174	44.16	697	8.90	300	50	12.0
5750	1.8	201	46.25	896	8.90	252	50	13.1
5750	1.9	182	46.10	777	8.90	275	50	12.5
5750	2.0	166	45.95	680	8.90	300	50	12.0
6000	1.8	193	48.07	874	8.90	252	50	13.1
6000	1.9	175	47.91	759	8.90	275	50	12.5
6000	2.0	159	47.75	663	8.90	300	50	12.0
6250	1.8	185	49.88	853	8.90	252	50	13.1
6250	1.9	168	49.71	741	8.90	275	50	12.5
6250	2.0	153	49.53	648	8.90	300	50	12.0
6500	1.8	178	51.69	834	8.90	252	50	13.1
6500	1.9	161	51.50	724	8.90	275	50	12.5
6500	2.0	147	51.32	633	8.90	300	50	12.0
6750	1.8	171	53.50	816	8.90	252	50	13.1
6750	1.9	155	53.30	708	8.90	275	50	12.5
6750	2.0	141	53.10	619	8.90	300	50	12.0
7000	1.8	165	55.30	798	8.90	252	50	13.1
7000	1.9	150	55.09	693	8.90	275	50	12.5
7000	2.0	136	54.87	606	8.90	300	50	12.0
7250	1.8	159	57.10	782	8.90	252	50	13.1
7250	1.9	144	56.87	678	8.90	275	50	12.5
7250	2.0	132	56.64	593	8.90	300	50	12.0
7500	1.8	154	58.89	766	8.90	252	50	13.1
7500	1.9	140	58.65	665	8.90	275	50	12.5
7500	2.0	127	58.41	581	8.90	300	50	12.0
7750	1.8	149	60.69	751	8.90	252	50	13.1
7750	1.9	135	60.43	652	8.90	275	50	12.5
7750	2.0	123	60.17	570	8.90	300	50	12.0
8000	1.8	144	62.47	736	8.90	252	50	13.1
8000	1.9	131	62.20	639	8.90	275	50	12.5
8000	2.0	119	61.93	559	8.90	300	50	12.0
8250	1.8	140	64.26	723	8.90	252	50	13.1
8250	1.9	127	63.97	627	8.90	275	50	12.5
8250	2.0	116	63.69	548	8.90	300	50	12.0
8500	1.8	136	66.04	709	8.90	252	50	13.1
8500	1.9	123	65.74	616	8.90	275	50	12.5
8500	2.0	112	65.44	538	8.90	300	50	12.0
8750	1.8	132	67.82	697	8.90	252	50	13.1
8750	1.9	120	67.50	605	8.90	275	50	12.5
8750	2.0	109	67.19	529	8.90	300	50	12.0
9000	1.8	128	69.59	684	8.90	252	50	13.1
9000	1.9	116	69.27	594	8.90	275	50	12.5
9000	2.0	106	68.94	519	8.90	300	50	12.0
9250	1.8	125	71.37	673	8.90	252	50	13.1
9250	1.9	113	71.02	584	8.90	275	50	12.5
9250	2.0	103	70.69	510	8.90	300	50	12.0
9500	1.8	122	73.14	661	8.90	252	50	13.1
9500	1.9	110	72.78	574	8.90	275	50	12.5
9500	2.0	101	72.43	502	8.90	300	50	12.0
9750	1.8	119	74.91	650	8.90	252	50	13.1
9750	1.9	107	74.53	564	8.90	275	50	12.5
9750	2.0	98	74.17	494	8.90	300	50	12.0
10000	1.8	116	76.67	640	8.90	252	50	13.1
10000	1.9	105	76.28	555	8.90	275	50	12.5
10000	2.0	95	75.90	486	8.90	300	50	12.0
10250	1.8	113	78.43	629	8.90	252	50	13.1
10250	1.9	102	78.03	546	8.90	275	50	12.5
10250	2.0	93	77.64	478	8.90	300	50	12.0
10500	1.8	110	80.19	620	8.90	252	50	13.1
10500	1.9	100	79.78	538	8.90	275	50	12.5
10500	2.0	91	79.37	470	8.90	300	50	12.0
10750	1.8	107	81.95	610	8.90	252	50	13.1
10750	1.9	97	81.52	529	8.90	275	50	12.5
10750	2.0	89	81.09	463	8.90	300	50	12.0
11000	1.8	105	83.71	601	8.90	252	50	13.1
11000	1.9	95	83.26	521	8.90	275	50	12.5
11000	2.0	87	82.82	456	8.90	300	50	12.0
11250	1.8	103	85.46	592	8.90	252	50	13.1
11250	1.9	93	85.00	514	8.90	275	50	12.5
11250	2.0	85	84.54	449	8.90	300	50	12.0
11500	1.8	100	87.21	583	8.90	252	50	13.1
11500	1.9	91	86.73	506	8.90	275	50	12.5
11500	2.0	83	86.26	442	8.90	300	50	12.0
11750	1.8	98	88.96	574	8.90	252	50	13.1
11750	1.9	89	88.47	499	8.90	275	50	12.5
11750	2.0	81	87.98	436	8.90	300	50	12.0
12000	1.8	96	90.71	566	8.90	252	50	13.1
12000	1.9	87	90.20	491	8.90	275	50	12.5
12000	2.0	80	89.70	430	8.90	300	50	12.0

You will notice that the Q of the Pi-L network does not go up as fast when frequency is lowered as it does with the pi network. Also, the Q remains the same for the pi-L for higher plate loads.

In both table 3 and table 4 the

numbers for ten meters are for 29.7 MHz, the highest frequency in the band. This is because you need to know minimum capacitance values to reach this frequency in a five-band transmitter.

effect of swr

A standing-wave ratio of 4:1 will affect the capacitance required at C1 by ±10%, and at C2 by ±25%. If the swr is caused by capacitive reactance, the tuning and loading capacitors are on the smaller side, and if the swr is the result of inductive reactance, the loading and tuning capacitors must be larger. Keep this in mind when selecting component values for a transmitter so you will be able to compensate for an antenna that is not exactly matched to your transmission line.

ssb and CW operation

Table 1 shows that for a given plate supply voltage the plate load impedance is inversely proportional to the input power level. That is, 1000 watts at 2800 volts represents 5000 ohms, while 2 kW at the same supply voltage is 2500 ohms. Pi network values for a Q of 12 for each of these impedances are shown below (capacitance in pF, inductance in μH):

f	R1	C1	L1	C2	Q
4.0 MHz	2500	191	9.18	1097	12.0
4.0 MHz	5000	95	17.38	534	12.0

Note that the required inductance values are quite different.

As an example of what you may expect under actual operating conditions, consider the 2-kW design above (9.18 μH inductor). At 3.5 MHz with 2 kW input, C1 is 252 pF, C2 is 1536 pF and Q is 13.8; not too bad. However, if the input power is reduced to 1000 watts at 3.5 MHz, C1 is 246 pF, C2 is 2287 pF, and Q reaches 27.0, *increasing* the circulating currents and heat losses in the network.

These figures point out the problems you can run into when you use the same operating voltage and same inductor at different power levels. Fortunately, there are several things which the designer can do to minimize these problems.

variable inductors

There are various variable rotary inductors available on the market which allow the operator to select the proper inductance for 1000 watts CW at the bottom of a band as well as 2000 watts PEP ssb at the top. When compared to fixed inductors, these variable units are fairly expensive, and require a turns counter, further increasing cost. However, they are available in various inductance and current-carrying abilities, so they encompass practically any design requirement.* Also, using a variable inductor eliminates the need for a bandswitch.

bandswitch

The primary purpose of the bandswitch is to change the tap on the tank-circuit inductor to one better suited for the band in use. However, there are several other important functions for the bandswitch:

1. Used to switch input networks to match the 50-ohm output of the exciter.

2. Changes taps on the second inductor in a pi-L network.

3. Sometimes used to switch in additional tuning/loading capacitance on the 80-meter band so smaller variable capacitors may be used in the circuit.

Since you may wish to use a bandswitch in your power amplifier because of these additional uses, the variable inductor may lose some of its appeal.†

tapping the inductor

In a novel approach to this problem a ten-position bandswitch has been used in

*A large selection of variable inductors is available from the E. F. Johnson Company, 1848 10th Avenue, SW, Waseca, Minnesota 56093.

†A good source for excellent high voltage bandswitches and variable capacitors is James Millen *Manufacturing Company*, 150 Exchange Street, Malden, Massachusetts 02148. Millen, unlike some manufacturers, is glad to sell to amateurs in unit quantities.

one design to select different amounts of inductance for the CW and ssb ends of the band.[4] However, the additional switch leads and the large number of inductor taps makes this approach seem impractical for the typical home builder.

Table 5 compares the performance of 1000- and 2000-watt transmitters, as well as a 2000-watt transmitter run at 1000 watts input. *In the latter case some* additional losses are evident, but they're hardly large enough to cause much excitement. The same comparison shows that the 2-kW transmitter with a Q of 12 at 4 MHz, has a Q of 16.2 at 3.0 MHz. However, considerably more capacitance is required at C1 and C2. The pi-L network will alleviate this problem to some extent as the Q of the pi-L does not increase as rapidly as frequency is lowered as it does with the straight pi network.

Since the 80-meter band is proportionally wider than any other high-frequency amateur band, there *is* some merit *in* using an extra bandswitch position for 80 meters.[5] While I have shown previously that this is not required, it would be beneficial because you could select the 75-meter inductor for 4 MHz and 2-kW input, with the 80-meter inductor chosen for 3.7 MHz and 1-kW input.

The primary advantage in such an arrangement would be the ability to add a second input network to match the exciter. Since the input networks have low Q (typically 2 to 3), they are quite broadband and are usually set to a frequency in the middle of the band. However, it would be literally impossible for the same input network to work equally well on both 3.5 and 4.0 MHz, so it would be desirable to have one for each end of the band. From a practical standpoint, this might not be necessary because most operators have ample drive on CW if they are able to push the final to 2-kW PEP on ssb.

table 5. Comparisons between a 1-kW transmitter, a 2-kW transmitter and the 2-kW transmitter operated at 1-kW input. A frequency of 4 MHz was used, but other frequencies from 3 to 30 MHz should produce comparable results. (These calculations are computer derived for comparative purposes and only approximate actual operating conditions.)

	1 kW	2 kW	1 kW on 2-kW transmitter	
Plate voltage	2800	2800	2800	volts
Plate current	357	714	357	mA
Plate load impedance	5000	2500	5000	ohms
Power input	1000	2000	1000	watts
Tube output (typical)	700	1400	700	watts
Power at antenna	672	1343	647	watts
Transmitter efficiency	67.2	67.1	64.7	percent
Network efficiency	96.0	95.9	92.5	percent
Lost in L1 (heat)	27.9	56.9	52.6	watts
Circuit Q	12	12	23.6	
Inductor Q (typical)	350	350	350	
Frequency	4.0	4.0	4.0	MHz
Antenna load	50	50	50	ohms
C1 tuning capacitor	95.5	191.0	187.8	pF
L1 inductor	17.38	9.18	9.18	μH
C2 loading capacitor	533.8	1096.9	1703.0	pF
C1 reactance	416.7	208.3	211.9	ohms
L1 reactance	436.9	230.7	230.7	ohms
C2 reactance	74.5	36.3	23.4	ohms
Current in C1	4.49	8.98	8.83	amps
Current in L1	4.73	9.29	8.93	amps
Current in C2	2.46	7.14	7.70	amps
Voltage across C1	2645.8	2645.8	2645.8	peak volts
Voltage across C2	259.2	366.5	254.4	peak volts
Voltage on antenna	183.3	259.1	179.9	rms volts
Current in antenna	3.67	5.18	3.60	amps

power supply voltage

Since, as I just mentioned, most exciters have more than ample drive for 1000-watts input on CW if they are capable of driving the final to 2000-watts PEP on ssb, it's desirable to include some sort of automatic swamping so the exciter can be run in a normal manner for both ssb and CW. Lowering the plate supply voltage on the final-amplifier tubes decreases the plate load impedance required for a given input power level, therefore requiring more drive to reach this input power level.

For example, if it takes 70 watts drive with 3000 volts on the plate to reach 2000-watts input, then, depending upon the tubes used, it would take 70 to 80 watts drive to reach 1000 watts input with a substantially lower plate supply voltage. At the same time, the voltage-current relationship has changed, lowering the plate load impedance to something much closer to that which would give a Q of 12 with the same inductor.

Also, the plate voltage must be lowered to retain the same Q with the same inductor at the same operating frequency. This voltage reduction can be determined from

$$E2 = 0.71\ (E1) \qquad (11)$$

where E1 is the original plate voltage for 2000 watts input and E2 is the lowered plate voltage for 1000 watts input. For example, a plate supply of 2800 volts for 2000 watts input must be changed to 2000 volts for 1000-watts input at the same operating frequency and circuit Q. Actually, on 3.5 MHz, this would be perhaps 1800 to 1900 volts to provide a Q of 12 at 3.5 MHz (1000 watts input) using a 2-kW transmitter designed for a Q of 12 at 4.0 MHz. However, it is unlikely that you could get 1000-watts input at this plate voltage, even with 100 watts drive on a cathode-driven grounded-grid amplifier.

tuning capacitance

Table 3 and **table 4** show that the C1 tuning capacitance becomes quite small

table 6. Large value at C1 and smaller inductor cause the Q on ten meters to rise very rapidly, especially when running the transmitter at a lower power input which requires 5000 ohms plate load impedance.

f (MHz)	R1	C1 (pF)	L1 (μH)	C2 (pF)	Q
29.7	2500	26	1.24	148	12.0
29.7	2500	32	1.00	210	15.0
29.7	2500	39	0.84	251	18.0
29.7	2500	45	0.72	300	21.0
29.7	2500	51	0.63	348	24.0
29.7	5000	26	1.24	234	24.0
29.7	5000	32	0.98	303	30.0
29.7	5000	45	0.70	437	42.0
29.7	5000	51	0.61	503	48.0

on 10 and 15 meters as the plate load impedance is raised. A typical 2000-watt transmitter might use 2800 volts on the plate, providing a plate load impedance of approximately 2500 ohms. This transmitter would require only 26 pF tuning capacitance to reach the top end of the 10-meter band.

Unfortunately, most modern rf power tubes designed for the 2000-watt level have output capacitances on the order of 10 pF – this leaves about 16 pF for tuning, including stray circuit capacitance.

If you study the various air-variable capacitors available you will find that it is virtually impossible to find a variable capacitor that will provide the necessary spacing for this operating voltage as well as tune the capacitance range needed for both 10 and 80 meters. Also, you must keep in mind that ±10% leeway should be provided to compensate for any swr on the transmission line.

As the plate load impedance increases, the situation becomes even more acute. A 1000-watt transmitter with a plate supply of 2800 volts has a plate load impedance of 5000 ohms – on ten meters this means the tuning capacitor C1 is a total of 13 pF. In this case you would probably have to delete C1 entirely from the circuit and let the capacitance of the power tube supply the necessary tuning capacitance. However, although this has been done, it is not practical.

Fortunately, there are several things you can do to help alleviate this situation. You can use a smaller capacitor and add fixed capacitance on 40 and 80 meters, or use two variable capacitors, switching in the larger one on the lower bands. The vacuum capacitor is another possibility because of its low minimum capacitance, often as low as 3 pF. You can also blunder ahead and use a too-large capacitor, allowing the Q to be higher than normal.

Oddly enough, each of these different techniques is currently being used in commercial amateur-band power amplifiers. The vacuum variable provides the best answer to this problem, but it is also the most expensive (by a wide margin). However, the vacuum variable has many advantages worth considering if you are more interested in performance than in total cost.

From table 6 you can see that the Q on ten meters goes up quite rapidly if too much capacitance is used at C1. One currently available commercial amplifier uses 2800 volts at 2-kW input (plate impedance, 2500 ohms). For ten meters this calls for an input capacitor of about 15 pF after the output capacitance of the tubes has been subtracted. However, this amplifier uses two 20-150 pF capacitors in parallel which are tuned in tandem with a geared arrangement. Thus, their minimum capacitance is about 40 pF, plus 10 pF added by the power tubes, providing a minimum input capacitance of more than 50 pF without any allowance for strays.

Table 6 shows that this gives a minimum Q of 24.0 at the top end of the ten-meter band (around 25.5 at the bottom end). If the amplifier were used at 1000-watts input, the Q would be nearly 48 at the top band edge and over 50 at the bottom!

This amplifier would obviously lose substantial power output in the form of heat in the tank inductor, and proper tuning would be very critical. It would also have to be retuned more often as frequency was changed.

This design is what I call the *blunder-ahead* method. In my mind, it would have been relatively simple for the manufacturer to use only one of the two tuning capacitors on 10, 15 and 20 meters, switching in the second tuning capacitor on 40 and 80.

Another manufacturer does precisely this. He uses a dual-section capacitor — half is used for the three upper bands and the other half is added in parallel on 40 and 80 meters. This provides normal Q for 2000 watts input on 10 meters. It still gives Q in excess of 20 with 1000-watts input, but that's really not too bad. This tuning system gives more than twice the *vernier* of the other system since the maximum capacitance on 20 meters, for example, is 120 pF. On the previous amplifier there is 300 pF available, even on 20 meters. The unit with the lower capacitance is far easier to tune on the upper three bands.

One other circuit trick which can be used quite successfully is to use a dual-section variable, placing the two sections in series rather than parallel. This reduces the minimum capacitance to 10 pF or less.

broadband power amplifier

Many operators need special frequencies outside the five amateur bands for MARS or other purposes, and need a power amplifier which can be tuned up at any frequency in the range from 3.0 to 30 MHz. Table 7 shows a pi-network design that gives continuous frequency coverage in five switch positions. A pi-L network for similar use is shown in table 8. The pi-L is more broadband for a given Q variation, and requires substantially less output capacitance. Both designs are for 2000 watts input with a 2800-volt plate supply, or 1000-watts input at 2000 volts.

component voltage ratings

To determine the peak voltage across C1 you *can* use the maximum dc plate voltage. This is not precisely correct, but it's close enough. Normally you would

increase the voltage by at least 30% when selecting a capacitor to prevent arcing if the tank circuit is not perfectly resonated, and to allow for some oxidation if you use an air variable.

There are several ways to determine peak voltage. If the power output is known at this point you can use eq. 12 to determine peak volatage:

$$E_{pk} = \sqrt{2PZ} \tag{12}$$

where E_{pk} is the peak rf voltage, P is output power and Z is plate load impedance. For example, in a 1-kW transmitter with 2800 volts on the plate, the peak voltage across C1 and L1 is 2646 volts. (The power output of class-B stages may be estimated at 70% of the input power as this gives some margin of protection and is suitable for this purpose.)

The peak voltage on C2 can also be figured in a similar manner, except that Z in eq. 12 is the antenna load impedance. Power output may be estimated at 65% of the input. For example, if the output power is 650 watts (for a 1-kW amplifier), and the antenna load is 50 ohms, this represents approximately 254 peak volts across C2. Thus, for a 1000-watt transmitter, a 350-volt, 365-pF broadcast receiver type capacitor could be used successfully. For 2000 watts input at 2800 volts, the peak voltage across C2 would be 367 volts, and the broadcast-tuning capacitor would be too marginal.

In the pi-L network the image impedance must be used when calculating the peak voltage across capacitor C2, and the voltage rating must be substantially higher than for the same capacitor in the pi network. For example, in a 1-kW transmitter, the peak voltage across C2 is about 635 volts: for a 2000-watt amplifier the peak voltage is about 895 volts.

component current ratings

The peak voltage across C1 has already been determined, but to find the current through C1, rms voltage is more useful. This can be found from eq. 13:

$$E_{rms} = \sqrt{PZ} \tag{13}$$

table 7. Pi-network component values for a broadband 3-30 MHz rf power amplifier matching a 50-ohm antenna load. This is accomplished in five bands: 3.0-5.0 MHz, 5.0-8.5 MHz, 8.5-14.4 MHz, 13.5-22.0 MHz and 20.0-30.0 MHz. The Q is set for a minimum of 12 at the top of each band. The 2500-ohm plate load impedance corresponds to a grounded-grid amplifier running 2000 watts at 2800 volts, or a 1000-watt amplifier with 2000 volts on the plate.

R1 OHMS	F MHZ	C1 PF	L1 UH	C2 PF	R2 OHMS	'Q' QUAL.
2500	3.0	433	7.34	2878	50	20.4
2500	3.5	317	7.34	2053	50	17.4
2500	4.0	242	7.34	1517	50	15.2
2500	5.0	153	7.34	878	50	12.0
2500	5.0	265	4.32	1764	50	20.8
2500	7.0	134	4.32	834	50	14.7
2500	7.3	123	4.32	755	50	14.1
2500	8.5	90	4.32	516	50	12.0
2500	8.5	155	2.55	1034	50	20.8
2500	14.0	56	2.55	327	50	12.4
2500	14.35	54	2.55	308	50	12.1
2500	14.4	53	2.55	305	50	12.0
2500	13.5	94	1.67	621	50	19.9
2500	21.0	38	1.67	225	50	12.6
2500	21.45	37	1.67	212	50	12.3
2500	22.0	35	1.67	199	50	12.0
2500	20.0	59	1.22	383	50	18.4
2500	28.0	30	1.22	176	50	15.0
2500	29.7	26	1.22	151	50	12.2
2500	30.0	25	1.22	146	50	12.0

where E_{rms} is the rms voltage, P is the output power and Z is the plate load impedance. In the previous example of the 1000-watt transmitter with a 2800-volt plate supply, the rms voltage across C1 is nearly 1870 volts.

To calculate the current through C1 you must first determine the reactance of C1 (eq. 14) and calculate its impedance (eq. 15). The current is found from eq. 16.

$$X_c = \frac{Z_p}{Q} \tag{14}$$

$$Z_{C1} = \sqrt{R^2 + X_C^2} \tag{15}$$

$$I = \frac{E_{rms}}{Z_{C1}} \tag{16}$$

However, since the resistance of a high-quality air-variable capacitor is very small, less than 1 ohm, for all practical purposes the impedance of the capcitor is equal to its reactance. Therefore, the current can be found from

$$I = \frac{E_{rms}}{X_{C1}} \tag{17}$$

As you can see in table 5, the current through C1 is much higher than you might think, with nearly 4.5 amperes flowing through C1 in the 1000-watt transmitter with 2800 volts on the plate. Most air variables and vacuum capacitors can handle this current easily, but you still must be careful when selecting fixed capacitors to pad the variables. Transmitting-type capacitors with high Q and good current-carrying capability are required (such as the Centralab 850 series).

The current through C2 can also be determined with **eq. 17**. However, when calculating the rms voltage across C2 the antenna load impedance must be used in **eq. 13**. Again, there is substantial current flowing through C2 — nearly 2.5 amperes in the 1000-watt transmitter.

For all practical purposes, the current through inductor L1 is equal to that through C1. It is actually a little higher, and the following formula is reasonably correct for class B:

$$I_{cc} = 1.05\, Q_o I_p \qquad (18)$$

where I_{cc} is the circulating current, Q_o is loaded circuit Q and I_p is the indicated plate current. **Eq. 18** is a close approximation that compares favorably with answers derived from using complex vector analysis of reactive components used in rf circuits at resonance.

inductor power loss

To determine heat losses in the inductor, it is necessary to know the rf resistance of the inductor. Then you can use **eq. 19** to find power loss.

$$P = I^2 r \qquad (19)$$

where *I is the circulating* current and r is the rf resistance.

To minimize these losses, the inductor should be silver plated, as should all leads to the bandswitch. Power losses on the order of 30 to 100 watts are not unusual, even with low standing-wave ratios. The use of tubing is encouraged, particularly on the higher frequencies to provide better unloaded Q.

table 8. Pi-L network component values for a broadband 3-30 MHz rf power amplifier matching a 50-ohm antenna load. This is accomplished in five bands: 3.0-5.0 MHz, 5.0-8.5 MHz, 8.5-14.4 MHz, 13.5-22.0 MHz and 20.0-30.0 MHz. The Q is set for a minimum of 12 at the top of each band. The 2500-ohm plate load impedance corresponds to a grounded-grid amplifier running 2000 watts at 2800 volts, or a 1000-watt amplifier with 2000 volts on the plate.

R1 OHMS	F MHZ	C1 PF.	L1 UH.	C2 PF.	L2 UH.	R3 OHMS	R2 OHMS	'Q' QUAL.
2500	3.0	388	9.00	1510	3.90	158	50	18.3
2500	3.5	292	9.00	1015	3.90	197	50	16.0
2500	4.0	228	9.00	717	3.90	242	50	14.4
2500	5.0	153	9.00	400	3.90	350	50	12.0
2500	5.0	237	5.29	935	2.29	154	50	18.6
2500	7.0	127	5.29	391	2.29	253	50	14.0
2500	7.3	118	5.29	351	2.29	270	50	13.5
2500	8.5	90	5.29	235	2.29	350	50	12.0
2500	8.5	139	3.12	549	1.35	154	50	18.6
2500	14.0	56	3.12	150	1.35	330	50	12.3
2500	14.35	54	3.12	141	1.35	345	50	12.1
2500	14.4	52	3.12	139	1.35	350	50	12.0
2500	13.5	85	2.04	323	.89	165	50	18.0
2500	21.0	38	2.04	103	.89	325	50	12.5
2500	21.45	36	2.04	97	.89	335	50	12.3
2500	22.0	35	2.04	91	.89	350	40	12.0
2500	20.0	53	1.50	192	.65	183	50	16.7
2500	28.0	29	1.50	80	.65	315	50	12.7
2500	29.7	26	1.50	68	.65	345	50	12.1
2500	30.0	25	1.50	67	.65	350	50	12.0

A suitable inductor for the L section of the pi-L network consists of two ***inches*** **of Air-Dux 1606T (6 turns-per-inch, no. 14, 2" diameter). It should be placed at right angles to the main pi inductor to avoid mutual inductance.**

frequency	number turns	approximate inductance
3.0-5.0 MHz	10.00	3.90 μH
5.0-8.5 MHz	6.75	2.25 μH
8.5-14.4 MHz	4.75	1.33 μH
13.5-22.0 MHz	3.50	0.83 μH
20.0-30.0 MHz	3.00	0.65 μH

It may come as a surprise to find that the conductivity of silver is only slightly superior to that of copper. In fact, a silver-plated coil is little more efficient than a *new* tank coil made of copper. Copper, however, oxidizes, and the outer rf-current-carrying layer becomes less effective. On the other hand, silver develops a form of silver sulfide on its outer surface which barely affects its conductivity. Over a period of years the silver-plated coil will retain most of its original conductivity.

safety

An rf choke should be used at the antenna output of any pi or pi-L net-

work. This choke should be large enough to blow the overload relay (or fuse) in the high-voltage power supply if the dc blocking capacitor should short out. This is the only backup protection you have to keep high dc voltage off the pi-network components if the blocking capacitor shorts out. This rf choke also keeps any dc component off the antenna.

RTTY and ssb

Many amateurs are interested in RTTY as well as CW and ssb. Since RTTY is essentially 100% key down, it's quite hard on the various components in the transmitter. On ssb, the typical duty factor is 30% to 50%, depending on how much ALC and other compression you use. Typically, however, the *average* circulating current in the network is perhaps one-third of that for key-down operation.

Table 5 shows that 2000-watts key-down gives comparable circulating currents to that of the same transmitter run at 1000-watts key-down with the same plate voltage and same inductor. This is due to the higher Q that is being used. Because of the lower duty cycle of ssb, running a 2000-watt transmitter key-down at 1000 watts for RTTY is three times as hard on the transmitter as running 2000-watts PEP! This is rather startling, and indicates why some rf power amplifiers should not be used on RTTY, although they are perfectly suitable for ssb at higher input power levels.

Conversely, it follows that if a manufacturer guarantees his unit to run indefinitely at 1000-watts key-down RTTY, that same transmitter should last forever at 2000-watts PEP ssb. Some manufacturers hedge if this specific question is posed to them.

summary

Using a 2000-watt rf power amplifier at the 1000-watt level for RTTY or CW poses certain inherent problems regarding heat and efficiency. High plate supply voltages raise the plate load impedance to the point where it may be difficult to get the minimum capacitance required for resonance on 10 and 15 meters.

When building a high-power final amplifier, consideration must be given to selecting components which will handle the voltage and currents encountered in the circuit. The formulas given in this article should make it relatively easy for the builder to predict what these voltages and circulating currents will be before he actually builds the amplifier.

Computer-derived tables provide much data for the builder, and clarify many design points only hinted at in previous articles. I hope that the information presented here will be of benefit to anyone who builds or buys a final rf power amplifier.

acknowledgements

Many people are interested in pi and pi-L networks, and have been of direct assistance. Providing particular assistance was Bob Sutherland, W6UOV, of EIMAC. I also spent a great deal of time reading articles written by George Grammer, W1DF, former technical editor of *QST*. His work in this field, and his series of three articles in *QST*[5] represent an outstanding contribution. Bill Craig, WB4FPK, was most helpful, as was Garey Barrell, K4OAH. Bill Carver, K6OLG, also provided stimulating comments.

The Computer Terminal Corporation of San Antonio, Texas, provided over 100 hours of computer terminal time which was invaluable in this project.

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ham radio

quick and easy

speaker driver module

A compact, high-output, low distortion audio module provides an easy short-cut in the design of solid-state receivers and mobile equipment

Jerry Vogt, WA2GCF, 182 Belmont Road, Rochester, New York 14612

First introduced by Bendix a few years ago, the BHA-0001, BHA-0002 and BHA-0004 hybrid solid-state audio amplifier modules provide two watts, fifteen watts and five watts of audio output, respectively, to drive a 3.2-ohm speaker from a low-level, high impedance audio source. The 1972 Allied Radio Industrial Catalog lists these as "amazing space-age microcircuit modules each containing a complete audio amplifier, employing the latest thick-film construction and requiring fewer external components than monolithic partial amplifiers. Its simplicity permits easy assembly in less than one hour, even for those with semi-technical skills, thus saving endless hours of needless drudgery." For those who have played with the typical complementary symmetry amplifier, as I have, you will be surprised to find out that you can place one of these devices down on a board with a few external capacitors and have your amplifier working within minutes without worrying about getting the bias and feedback figured out for that long string of dc-coupled transistors.

I decided to try the BHA-0004, which provides a good amount of audio output (five watts) for a mobile rig. The rather high price, ($18.80 in single lots) was discouraging when I first looked at the catalog, but you only live *once.* Built as a module separate from the rest of the

receiver, you can use it over and over on new projects.

I started thinking about possible uses for the unit. The list seemed almost endless. The circuit can be terminated at the input with a 3-ohm resistor and used as an audio booster for mobile gear with flea-power output stages. It can be used to drive one car speaker with many different radios. Each radio is fed to the module input through low-resistance pads in place of the existing radio's speaker. You can use it to drive an electronic siren or a speaker mounted under the hood of the car, or as the basis for an intercom system. Two of them make a stereo amplifier for an fm tuner. I have built quite a few fm receivers, using the Sprague ULN-2111A IC limiter/detector and I am now starting to use this audio amplifier as a standard external module for my new receivers. The level from one of the fm detector ICs is perfect for driving these speaker amplifiers.

the circuit

The equivalent circuit for the device is shown in **fig.** 1. It is very similar to a circuit once promoted by Motorola for use with discrete transistors, using the 2N4918 and 2N4921 complementary output transistors. Basically, the internal circuit of the BHA-0004 consists of dc-coupled transistors. Output transistors, Q4 and Q6, operate as class-B symmetrical amplifiers, driven by Q3 and Q5. C1 suppresses high-frequency oscillations. R9 and R10 limit the output current that can be drawn through Q4 and Q6 from the external load. CR1 and CR2, in conjunction with R8, provide the voltage drop required between the bases of Q3 and Q5 to prevent cross-over distortion. If the bases of Q3 and Q5 were tied directly together, the emitter-base forward turn-on voltage of the two transistors would provide about 0.7 volt of reverse bias to the transistors, causing distortion, especially at low signal levels.

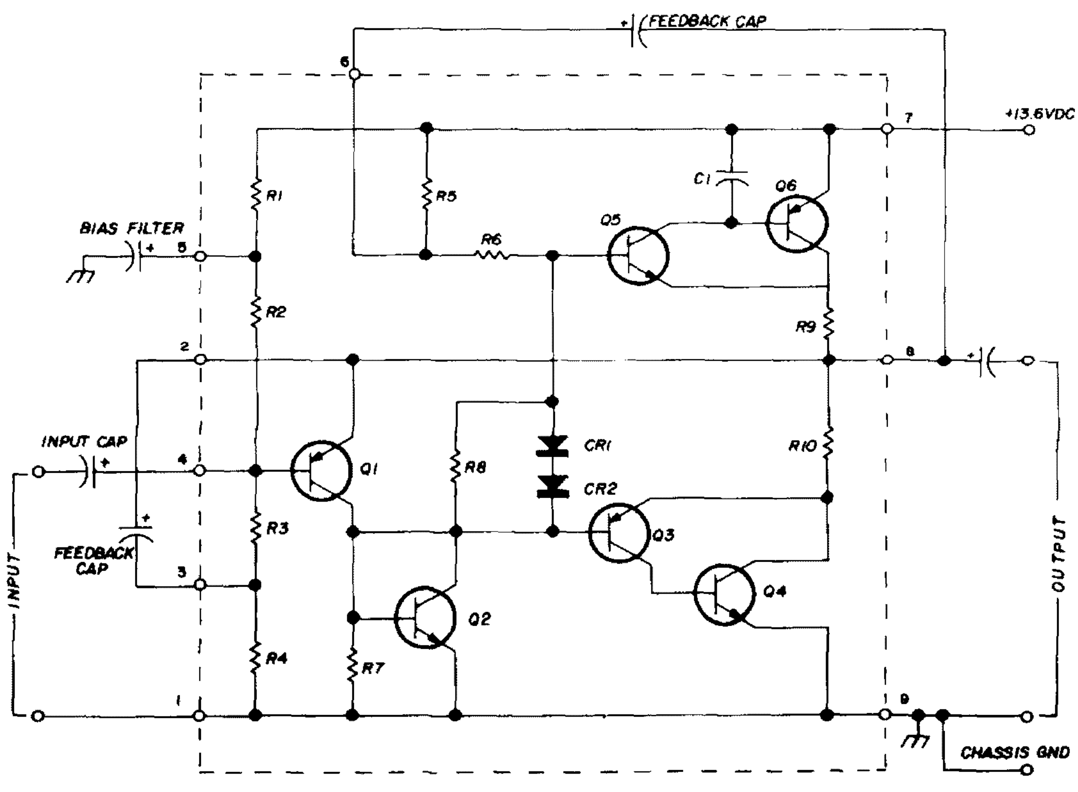

fig. 1. Equivalent internal circuit of the BHA-0004 audio amplifier, with external components added for understanding.

R5 and R6 provide a dc voltage to the biasing circuit for Q5 and Q3.

The easiest way to visualize the coupling from Q2 to the Q3 and Q5 pair is to see Q2 as a variable resistance in the bottom leg of a voltage divider consisting of R5-R6-CR1-CR2-Q2. In this way, Q2 modulates the bias applied to both Q3 and Q5 and drives them. It is at this point, where Q2 drives complementary transistors Q3 and Q5 (opposite polarities), that the single-ended signal is split to push-pull for the output stage. R8 provides compensation over the turn-on curve of CR1 and CR2. C2 suppresses parasitic oscillations. R7 acts as a load for Q1, the input transistor, which is directly coupled to the base of Q2. R1 through R4 provide dc biasing for the entire string of transistors by establishing the current flow through Q1 at idle. A tap in the circuit between R1 and R2 allows an external capacitor to be connected to filter or decouple the bias circuit.

A tap between R3 and R4 and one between R5 and R6 allow external capacitors to be connected from the output back to the input of various stages. This feedback minimizes distortion and levels the frequency response. The input is ac coupled to the base of Q1, and the output is capacitively coupled to the load, usually a loudspeaker. The circuit is optimized for operation from the normal car battery voltage of 13.6 Vdc. Preset idle current and center voltage provides ideal operation over a wide range of load conditions. Notice the benefit of this circuit in not requiring an output transformer.

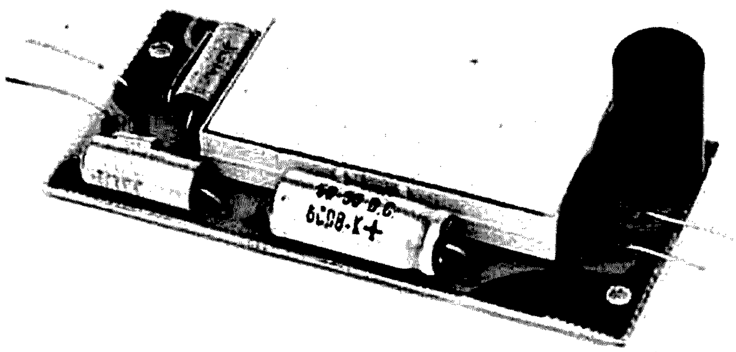

Photo of the completed audio module on the printed-circuit board.

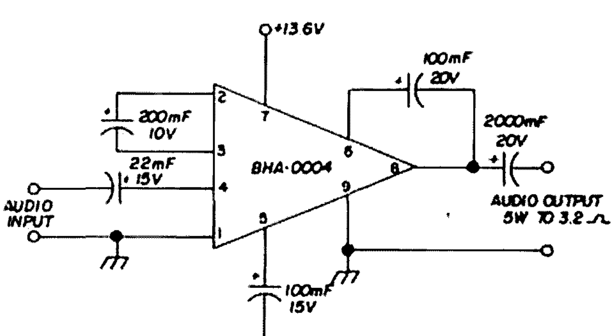

fig. 2. Test circuit recommended by the manufacturer.

external circuitry

The external connections recommended by the manufacturer are shown in fig. 2. The values given were selected for hi-fi operation, though, and are undesirable for amateur use in communications circuits. If you want to operate the module for music reproduction, however, that is the route to take. The components shown also allow you to obtain full power output, but that was not a requirement in the design which follows. Specifications for the unit in this "hi-fi" setup are 5 watts output over the range 25 Hz to 15 kHz with an input of 20-mV rms. Distortion should be 1% or less.

Although the manufacturer states that

no heat sink is required for full-power operation at room temperature, I think it would be a good idea to cement a few aluminum fins to the ceramic top of the module for protection. Remember that a replacement is expensive if you exceed the limits a little bit. The fine print with the device tells you to consider heat-sinking for operation over two watts at elevated temperatures. The thing to remember is that the unit itself generates some heat, and that heat will raise the ambient temperature if the device is operated in a package which allows no air circulation – a common method of packaging solid-state mobile gear.

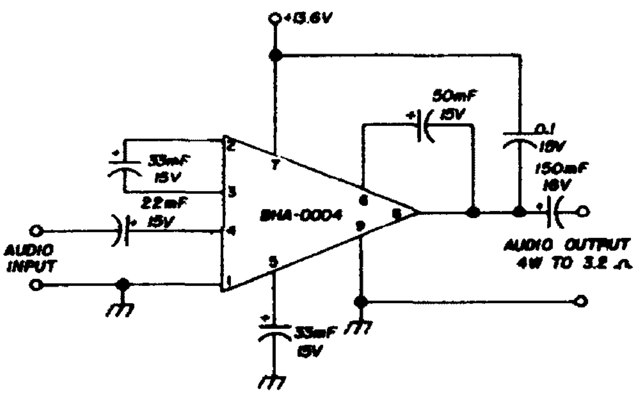

fig. 3. The modified circuit for compact packaging and communications work.

construction

The circuit as modified for use is shown in fig. 3. There are three reasons for the modifications: to reduce the size of the assembled circuit board, to restrict the frequency response to the communications range and to save the cost of unnecessary expensive parts. Basically, the values of all of the capacitors were reduced. An added component, a 0.1-mF capacitor from the B+ line to the output line prevents high-frequency oscillations found in one of the breadboard models.

With the circuit shown, the sensitivity is not quite as great as with the manufacturer's recommended circuit; however, it was felt to be adequate, especially in light of the high output normally available from the fm detectors in current use in IC receivers. The modified circuit provides four watts output over the 350- to 3500-Hz range with an input of 750-mV rms. No heat sink is required with this installation. The size of the circuit board, when assembled with the components in fig. 3, is approximately 2 1/8 x 3 1/4 x 1-inch high. This size works out nicely for mounting in many of the cases available to the home builder.

There are really no secrets or tricks used in construction of the assembled circuit board.* Fig. 4 shows the basic layout I used. The board layout was based on readily available parts, and spacing was set up to allow substitution if necessary. I replaced the resistor from pin 2 of the module to the 33-mF capacitor with a plain jumper. I found the resistor unnecessary. The only part at all out of the ordinary is the 150-mF capacitor,

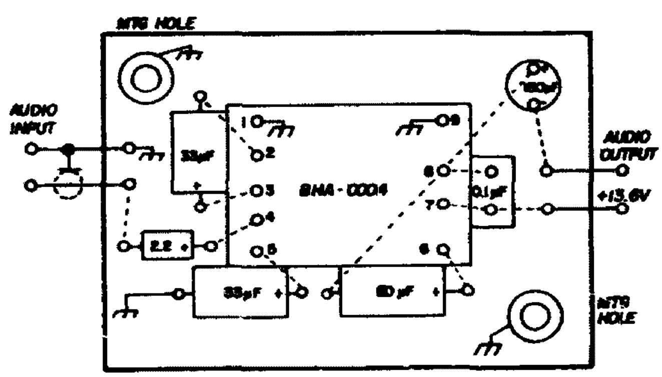

fig. 4. Suggested layout for a printed-circuit board (not to exact scale). The layout is very flexible and can be modified to suit the parts on hand.

which is a vertically mounted type which happened to be available. A more standard tantalum type (for small size) can be mounted in its place, without increasing the size of the board, by mounting it vertically with one lead folded over the side of the capacitor.

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*In conjunction with this article, a limited supply of completely assembled and tested assemblies are being made available for $19.50 postpaid. Contact Hamtronics Inc., 182 Belmont Road, Rochester, New York 14612. Please include remittance with order and allow two to three weeks for delivery.

three-band high-frequency log-periodic antennas

These three-band wire-beam log periodics for 20, 15 and 10 meters are inexpensive, easy to build and provide excellent performance

G.E. Smith, W4AEO, 1816 Brevard Place, Camden, South Carolina 29020

Log-periodic antennas offer a number of operating advantages to the amateur who wants consistent contacts over long distances. Although there are several commercial rotatable high-frequency log periodics available on the market, they are large and complex, and home construction of a rotatable L-P is impractical for the average amateur. However, the same performance can be obtained from a light-weight wire log periodic which is fixed in one direction.

The log periodics described in this article are wire beams, so they are low cost and easy to build. The log-periodic antenna shown in fig. 1 covers the frequency range from 14 to 30 MHz and can be used on the 10-, 15- and 20-meter bands. It can be erected in a 40- by 50-foot space and provides a minimum of 8-dB forward gain, a front-to-back ratio of 15 dB or more, and has low swr that is constant over each of the three amateur bands.

Although log periodics can be designed to cover a 10:1 frequency range, they are quite large. For this reason the antennas discussed in this article are limited to fixed, non-rotatable types for 20, 15 and 10 meters.

Since all details of log-periodic theory and design have been covered in previous articles, this data will not be repeated

here.[1,2,3] I have put up a number of fixed log periodics in various directions, and for different frequency ranges, including L-Ps for 20, 15 and 10; 20 and 15; 15 and 10, and a big brute for 40, 20 and 15. All the L-Ps installed so far have provided excellent on-the-air performance.

At my station these antennas are suspended from tall pine and cedar trees, with the elements 45 to 50 feet above ground. All were originally beamed south so I could evaluate their performance rapidly with the rather consistent band openings I have to South and Central America.

The log-periodic antenna illustrated in fig. 1 has an apex angle of approximately 36° (α=18°). If you want higher forward gain, and if space is available, the design in **fig. 2** has a minimum gain of 10 dB for each of the three bands. However, its overall length is 100 feet. This antenna has been in use at my station for the past year, and has done an excellent job.

When operating on 20 meters, using an ordinary dipole (at the same height as the log periodics), reports from South and Central America average S8 to S9. When I switch to the log periodic the signal reports usually improve to 20 dB over S9. In most cases the S-meter in my receiver confirms this.

Although these reports seem to indicate gain greater than 10 dB, when compared to the dipole I use as a standard, some of the apparent gain is probably due to the lower radiation angle of the log periodic. Also, the theoretical gain of a log periodic is the result of line-of-sight tests on vhf and uhf antenna ranges, so they are not directly translatable to high-frequency performance.

If you check the specification sheets for commercial log-periodic antennas, you will find that the manufacturers rate their 12- and 13-element log periodics at 10 to 13.5 dB *over average soil conditions.* Front-to-back ratios are rated from 14 to 16 dB.

The lower radiation angle of the log periodic always results in higher performance than that predicted by theory, particularly on 20 meters. And, the longer the DX path, the greater the difference when compared with a dipole.

Operational tests on 15 and 10 meters have not been as outstanding as those on 20 meters, but most reports give at least a 10 dB advantage to the log periodic.

Reports off the back of the beam generally show a front-to-back ratio of at least 15 dB (also confirmed by my receiver S-meter). The front-to-back ratio is generally best on 20 meters, and slightly less on 15. The conditions on 10 meters have been too erratic to make good front-to-back signal-level comparisons.

One of the big operating advantages of the log periodic is the apparent diversity effect on receive. This is particularly noticeable during conditions of severe fading. Even signals coming in from the back of the antenna often have less fading when compared to a dipole. Evidently the large size (large capture area) of the log periodic provides this effect.

Since the log periodic is a broadband antenna it is well suited for operating on any frequency within the amateur bands it is designed for. The swr is low and nearly constant over the entire length of each band. Also, because of its broadband characteristics, there are no critical element or impedance-matching adjustments necessary after you put it up.

theory

According to log-periodic theory, the longest rear element must be at least 5% longer than one-half wavelength at the lowest desired operating frequency. For example, if the lowest operating frequency is 14.0 MHz ($\frac{1}{2}\lambda$=33.4 feet), the rear element must be not less than 35 feet long (33.4 feet + 5%). This element would resonate at about 13.3 MHz.

The shortest front element should be 45% to 50% shorter than one-half wave-

length at the highest operating frequency. With 29.7 MHz as the upper frequency limit of the antenna, the front element should be resonant at 44.55 MHz mini-

would only require a space about 25 feet wide by 35 feet long. You could even build a rotary log periodic for 15 and 10 meters on a 25-foot boom.

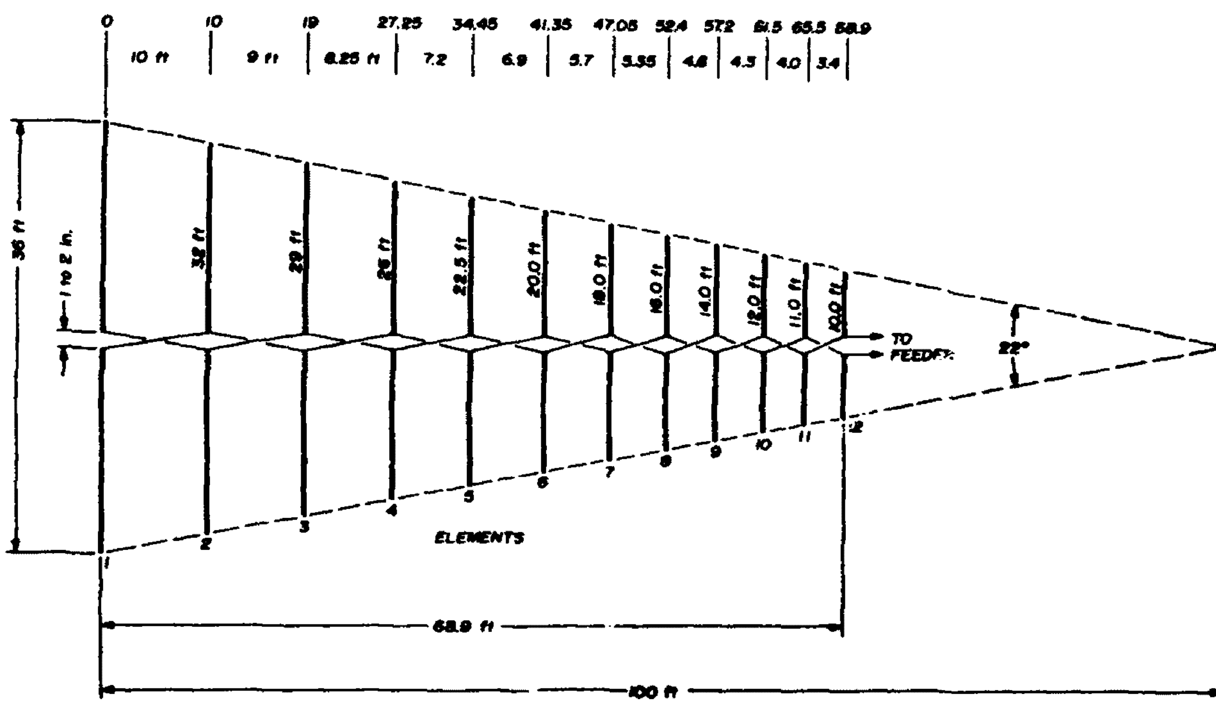

fig. 1. Three-band log-periodic antenna for 20, 15 and 10 meters provides approximately 10 dB forward gain.

mum (29.7 MHz + 50%).

For these reasons, a log periodic designed for 20, 15 and 10 meters should have a low-frequency cutoff of 12 to 13 MHz; high-frequency cutoff should be at least 45 MHz.

The log-periodic antennas shown in figs. 1 and 2, since they have the same number of elements, cost about the same, $35.00. However, this does not include support masts or feedline. This is not too bad for an antenna which provides 10 dB gain and performs equally well over each of the three amateur bands, 20 15 and 10.

If you don't have the space for one of these large antennas, you can reduce the size by eliminating one of the bands. For example, if 20-meter operation is not required, the three rearmost elements can be deleted. This leaves a 9-element log periodic that performs admirably on 15 and 10. The smaller, two-band antenna

If 10-meter operation is not required, you can remove the three front elements, leaving 9 active elements for operation on 20 and 15. This reduces the length of the antenna by about 6½ feet.

construction

Since I use tall trees around my house to support the log periodics, weight must be kept to a minimum to gain maximum height. For the antenna elements, I use no. 15 aluminum electric fence wire which is available from Sears (catalog no. 13K22065). This wire is very inexpensive at $8.70 for ¼ mile of wire and is extremely light weight and easy to work. It has good strength and should also be suitable for rhombics and other long-wire antennas.

Connections to the aluminum wire are made by winding no. 16 or no. 18 tinned copper wire around the aluminum wire for about one inch. The junction is then

covered with plastic electricians tape to keep out the rain and minimize electrolysis between the two dissimilar metals.

All the center insulators used in my

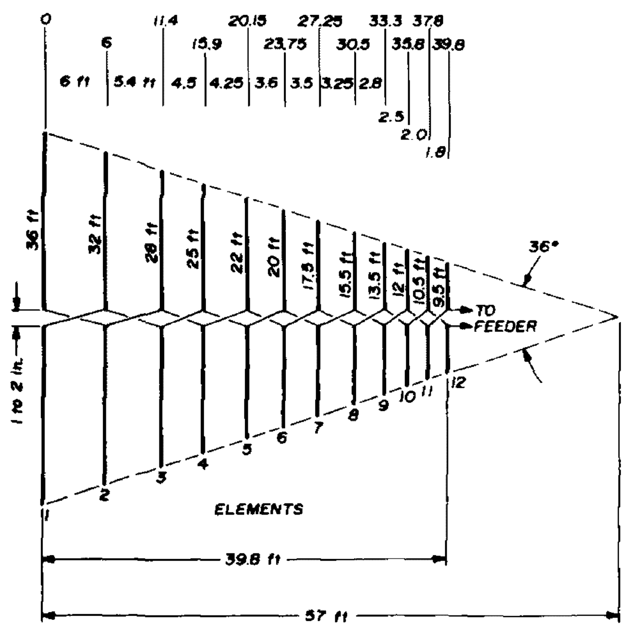

fig. 2. This three-band log periodic provides nearly the same performance as the design shown in fig. 1, but requires slightly less space.

log periodic were made from Lucite or Plexiglass sheet, 1/8- to 1/4-inch thick. After cutting and drilling, the center insulators cost about 20 cents each. These insulators are also used as spacers and stringers for the open-wire feeder which runs down the center of the log periodic. For the three-band log-periodics shown in **fig. 1** and **2** you will need 12 center insulators.

The end insulators are made from monofilament *fishing line* (40 to 50 lb. test). At the rear element, however, if you use two rear masts, Isolantite antenna insulators should be used at the ends of the elements because the strain at this point is quite high, and may exceed the rating of the monofilament.

The monofilament apparently provides more than adequate insulation for 1000-watt transmitters. YV5DLT has advised that he has experienced no breakdown problems when using his SB-220 linear at full output. The monofilament I used is Sears Catalog number 6KV32232 (40 lb. test), which is priced at $1.88 for a 325-yard spool.

installation

The log periodic is suspended from the center with 3/16-inch nylon line (A in **fig. 4**) and two side catenary lines (C in **fig. 4**). The 3/16-inch, 800 pound test nylon line used for the A line carries most of the load and strain of the antenna, including the open-wire feeder and the center insulators.

Before installing the antenna, string the center nylon line through the ¼-inch hole at the top of each of the 12 center insulators. After the insulators are on the line, stretch the line between two posts about 60 or 110 feet apart (depending which log periodic you are building). The line should be at shoulder height so it's easy to work on.

The first center insulator will support the longest element as well as the rear end of the center feeder. A knot is tied in the A line just in front of the insulator to keep it from slipping forward on the line. Make sure the other 11 insulators are on the other side of the knot.

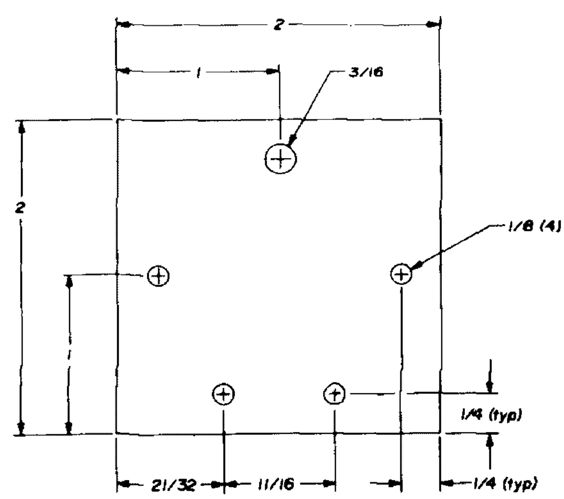

fig. 3. Insulators used in the construction of the log-periodic wire beams. Material is ¼" Lucite or Plexiglass.

Wrap several layers of masking tape around the nylon line to the rear of the first insulator. Leave a little space between the tape and the insulator so it hangs freely from the nylon line.

Now, using a steel tape, measure the spacing to the second insulator. Secure this insulator in place with several layers of tape around the line on each side of the insulator. Be sure to leave enough space between the tape layers so the tape or plastic tape, which often loosens up. The masking tape hardens and keeps the insulators in their correct position.

After the center insulators have been installed, assemble the parallel open-wire feeder by threading the two stranded

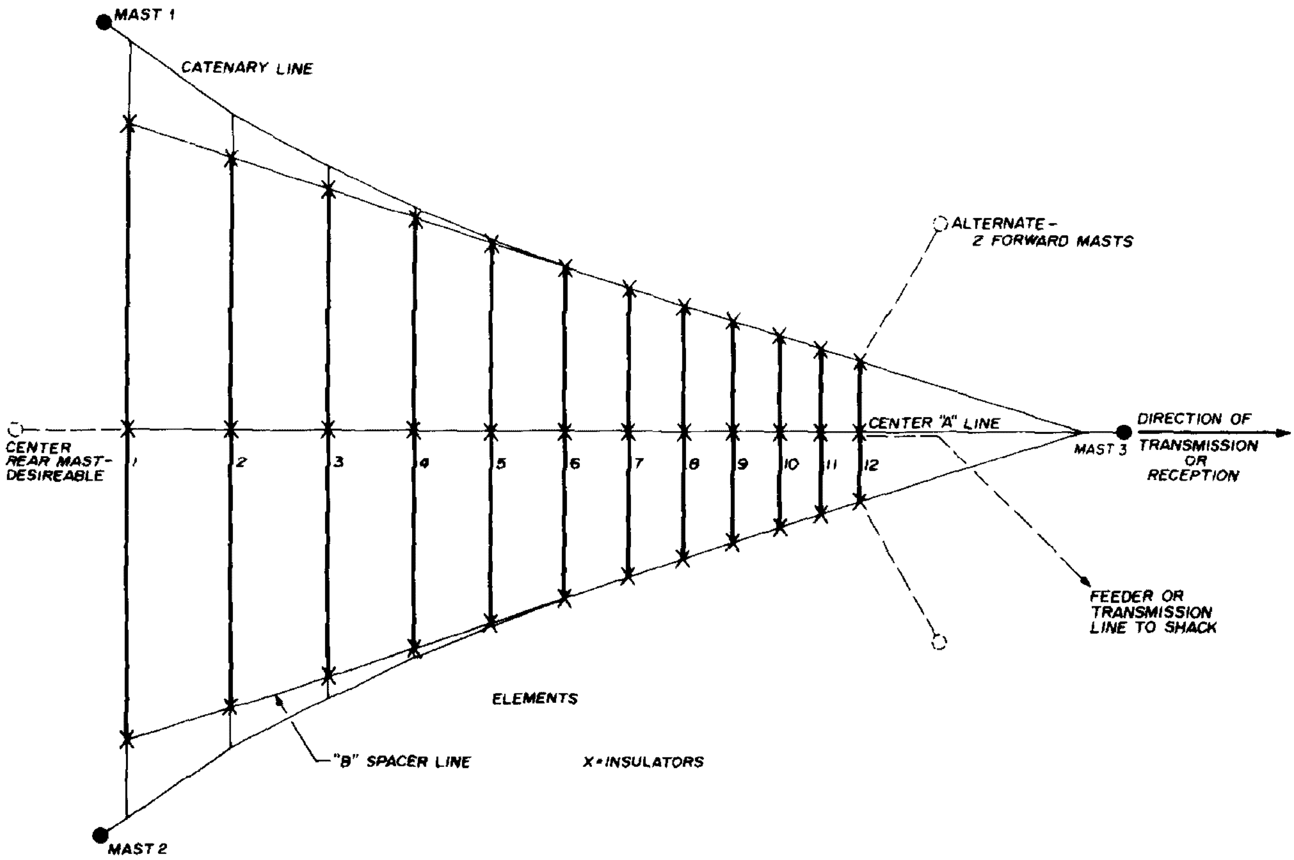

fig. 4. The log-period wire beams are supported by three nylon lines: the center A line and the two C catenary side lines.

insulator hangs freely on the center support line.

Continue along the center support line, measuring element spacing and installing center insulators, until all 12 insulators are correctly spaced and secured to the line. When spacing the insulators along the line it's a good idea to check the total spacing to make sure that no additive errors occur as spacing progresses. The total distance from the first insulator to the last should be 39.8 feet for the log periodic in fig. 2. For the larger antenna in fig. 1 this distance is 68.9 feet.

I have found that masking tape will stand the weather better than friction wires (7/24 or equivalent) through the two number-2 holes in the center insulators. The parallel feeder wires are secured to the insulators with a few turns of no. 18 wire as shown in fig. 5.

The spacing of the center feeder does not appear to be critical. I have used spacings from 3/4 to 2 inches on different log periodics. Some of the commercial vhf-tv log periodics have center spacing up to 5 inches. No doubt this spacing could be used on high-frequency log periodic antennas, but the larger spacing would require more Lucite for the center insulators, and this would increase both cost and weight.

When the center feeder is in place, cut

the 12 elements from a length of aluminum electric-fence wire using the dimensions shown in the illustrations. Make the elements slightly longer, so there is several inches of wire for attaching the monofilament end inculators, and at least 8

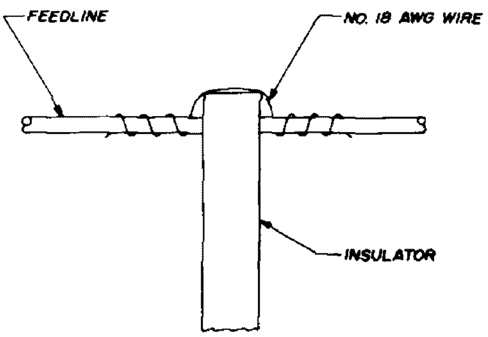

fig. 5. Lucite insulators are secured to the center feedline with a few twists of no. 18 wire.

inches for attaching the center insulators and connecting to the center feeder. Every other element is transposed as shown in **figs. 1** and **2.**

Attach the 12 elements to the center insulators, starting at one end of the antenna and working toward the other. Leave the ends at the center disconnected for the time being. After all the elements are attached to the center insulators, install the monofilament end insulators.

The two B lines are simply longitudinal spacers to keep the elements parallel during assembly and at right angles to the center feeder. These lines can be 1/8-inch or smaller since they carry no load (number 18 nylon twine, 165 lb. test, is satisfactory, and inexpensive at $1.85 for a 500-foot spool).

Next, the elements are attached to the *two C bridles or catenary lines* which take most of the side load. For the C lines, 1/8-inch nylon (375 to 500 lb. test) is suggested. This can be purchased at marine or hardware stores for about 3 cents per foot.

The center A line and the two C catenaries should be stretched tight, about 6 feet off the ground. When the catenaries are stretched into place they will appear as a large V, with the apex aimed in the desired operating direction. The A line should pass through the center of the V, bisecting it equilaterally.

By suspending the complete antenna between the same supports that will be used *in* the final installation, but at six feet above the ground, it is easy to adjust the tension of the elements from the ground.

The distance from the shortest element to the apex of the V should be 17.2 feet (fig. 1) or 31.1 feet (fig. 2). Less than this will allow the front element to sag too much.

Attach each of the elements to the catenaries with nylon twine, working from the shortest element to the longest. Use temporary knots, because it may be necessary to adjust the tension after all the elements are installed. Note that the six front elements usually fill the space between the B and C lines where the B *line is adjacent* to the C line.

Starting at element number 7, the C lines will require more and more separation to provide sufficient tension on the longer rear elements.

At this point it may be necessary to adjust the spacing between the end insulators and the C lines so there is as little element sag as possible, but don't put too much strain on the nylon support lines. There will also be some fore and aft sag of the center A line due to the weight of the feeder, insulators and wire elements, but the antenna should now be starting to take shape.

center feed

The center feedline to each of the elements of the log periodic must be transposed as shown in figs. 1 and 2. I have tried two methods of doing this. On the antennas shown here each feeder is transposed 180° between each of the elements. This is the system usually used in the schematic representations of the log periodic.

With this method of feeding power to the elements, insulated wire must be used for the feeder. With high power, you might have problems with insulation

breakdown. Bare wire can be used for the feeders, but insulated transposition blocks are required between elements, adding both weight and cost.

The second center feed method uses an open wire parallel feeder with criss-crossing wires to each of the elements as shown in fig. 6. This feed system is easier to build, and presents a neater appearance.

feedline

Most of the rotatable vhf and uhf log-periodic antennas proviously described in the amateur radio magazines have used 50- or 72-ohm coaxial feedlines.[4,5,6] However, a coaxial feedline is not suitable for the high-frequency log periodics described here because the cable is much too heavy. For these antennas, a light-weight feeder is required.

Normally, the log periodic is fed from the front (short-element) end. The input impedance at this point is about 30 to 35 ohms, as measured with an Omega Antenna Noise Bridge.* I checked several different log-periodics with the Noise Bridge, and all fell into the 30- to 35-ohm range.

However, if the open center feed is extended to the apex, the input impedance increases to approximately 100 to 300 ohms.[8] The open center feed operates as an impedance transformer, and at a point that is an odd number of wavelengths from the active elements on 20, 15 and 10 (20 meters, element 2; 15 meters, element 5; 10 meters, element 8) the input impedance remains fairly constant over each of the amateur bands. This point is within several feet of the apex.

Since the input impedance of the antenna depends upon feed point location, several possible transmission lines may be used. Since the input impedance at the front element is quite low, one of the best methods of feeding the antenna is with tuned open-wire feeders, with an antenna tuning unit between the coaxial output of the transmitter and the open-wire feed-line.

Although a coaxial feedline adds a great deal of weight to the antenna, and results in a sagging log periodic, it can be connected directly, through a balun, to the front-element feed point. Tests here, with RG-8/U coaxial cable, indicate a fairly good match on 20, 15 and 10. The standing-wave ratio on the 14-MHz band ranges from 1.1:1 to 1.3:1 from one end of the band to the other. On 21-MHz, the swr varies from 1.3:1 to 1.7:1, and on 28 MHz, the swr is from 2.0:1 up to 2.5:1.

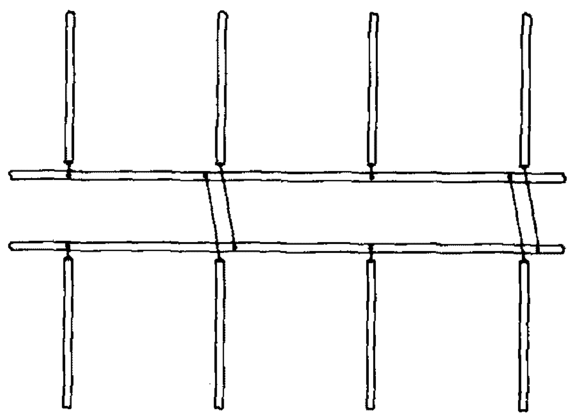

fig. 6. The required feedline transposition is most easily accomplished *with criss-crossing* wires to each of the antenna elements.

The swr on 10 meters is somewhat higher than that on the two lower bands, but it is still within tolerable limits, and on-the-air tests on 10 meters indicate very good performance.

At my station neither the tuned open-wire feeder nor the coaxial cable were suitable for a permanent installation. Since I use trees to support the antenna, the weight of the coax cable caused the antenna to sag too much, and valuable height was lost. Also, several of these antennas are several hundred feet from the station, so the cost of coaxial feedline is prohibitive.

The long length of the feedline makes open-wire feeder impractical because of the large number of spacers to be in-

*When using the Antenna Noise Bridge, the *frequency* dip normally exhibited by a sharply tuned antenna is completely absent with a log periodic because of its broad-band operation.

stalled, and the amount of work required to install the transmission line.

For these reasons I tried 72- and 300-ohm tv twin-lead. I tried the 72-ohm twin-lead first, connecting it to the transmitter through a short section of coaxial cable. For minimum swr (at the transmitter) I had to prune the length of the 72-ohm feeder – by removing short lengths (about 1/8-wave at 28 MHz), and making swr measurements, I arrived at a feeder length which provided fairly low swr on each of the three bands. A 1:1 balun between the twin-lead and the coax input didn't appear to make any difference.

When I tried out the 300-ohm twin-lead, connected near the apex, I used a 4:1 balun transformer between the twin-lead and the coax to my transmitter. This system worked quite well, and provided good performance on all bands. Although tv-type twin-lead will not handle a kilowatt, it is adequate for the 250 watts which I use. For higher power installations, transmitting-type 300-ohm twin-lead is available.

With the 300-ohm twin-lead feedline, the swr on 14.0 MHz was measured at 1.7:1, dropping to 1.5:1 at 14.2 MHz and 1.3:1 at 14.35 MHz. On 21.0 MHz the swr was 2.2:1, increasing to 2.5:1 at both 21.2 and 21.45 MHz. On ten meters, the swr was 2.2:1 at 28.0 MHz, dropping to 1.9:1 at 28.5 MHz, increasing to 2.1:1 at 29.0 MHz, and dropping again to 1.9:1 at 29.5 MHz. When plotted on a graph, these swr figures result in pretty flat performance over each of the three amateur bands.

summary

Since the forward lobe of the log periodic is generally broader than that of a Yagi, it is quite suitable as a fixed, non-rotatable, gain antenna. When my antenna farm is completed, I will be able to cover the United States and several continents with six three-band log-periodic antennas. Six dpdt relays will be used to connect the desired 300-ohm feeder to a 4:1 balun which is connected to the coaxial transmission line to the transmitter.

These light-weight log periodics have been very durable. One has been up for a year, with absolutely no trouble. Three were up last winter and withstood two bad ice storms; they sagged a bit with the ice load, but as soon as the ice melted, they returned to their normal height. They have also withstood a couple of twisters which passed a block away, snapping a number of tall pines.

If you like to build and test antennas, or are looking for DX and consistent contacts in a certain direction, I highly recommend the light-weight log periodics described here. At the present time I am working on two side-by-side log periodics pointing in the same direction. This should increase the gain by about 3 dB, to 13 dB for the two-antenna system.

I want to thank all those who have been helpful in giving reports and running tests on the log periodics, especially the Central and South American operators, for their patience and accurate reports. Special thanks goes to YV5DLT in Caracas for the many tests over the past year, and the nearly daily schedules during the design and testing of these antennas.

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ham radio

RTTY distortion:

causes and cures

Anomalies peculiar to transmitted and received RTTY pulses are analyzed, with suggested improvements to reduce printout errors

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Many amateurs who pursue RTTY go to great lengths with fsk demodulators, keyers, and related equipment to get near-perfect printouts. Nearly every innovation is discussed, tried, and eventually designed into RTTY stations. Pulse distortion, however, seems to be rarely appreciated and often turns out to be the culprit causing many unnecessary printout errors.

When transmitting, it is the responsiblity of the RTTY amateur to send a clean signal with near-zero distortion. Frequency-shift circuits must be designed with regard to this requirement; and key-boards, TDs, and other transmitting devices require careful adjustment.

Distortion must also be considered in receiving circuits to minimize errors. This requirement applies not only to the demodulator but also to the dc loop circuit and the adjustment of the teleprinter and reperforator. This article defines common distortion terminology and discusses some of its causes, remedies, and measurement techniques.

definitions

Pulse distortion occurs in two ways – by *bias* and *end distortion.* Each can have two polarities: mark and space. As shown in **fig. 1**, when a transmitted or received RTTY signal is compared with a perfect signal carrying the same information and referenced to the space-going

edge of the start pulse, the following is evident:

a. *Bias* affects the mark-going pulse edges.

b. *End distortion* affects the space-going pulse edges.

Each type of pulse distortion is measured in percent and defined over the range 0-100 (**fig. 2**). Note that the signal completely disappears as bias distortion approaches 100 percent.

Distortion, D, can be expressed as

$$D = 100B(\Delta T)$$

where B is the baud rate in units per second and ΔT is the time error in seconds of the edge in question as measured from the correct edge location.

sources of distortion

Any circuit condition that imparts a delay to RTTY pulses acting on one polarity more than another will generate bias distortion. End distortion, in pure form, is rare since its effect on a signal selectively leaves the start pulse intact and is thus usually accompanied by equal or greater parts of bias distortion. Here are some common examples.

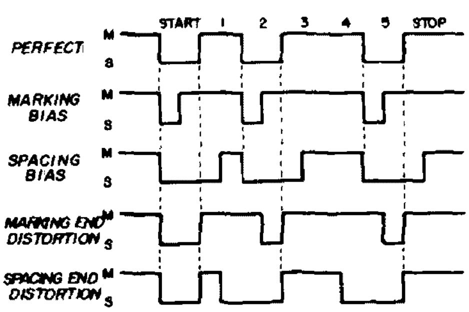

fig. 1. Basic distortion types on the letter *F*.

Inadequate open-loop voltage. When a teleprinter loop includes a large series inductance such as a selector magnet, the current response, which is the parameter that relates to the force applied to the selector armature, is distorted (**fig. 3**). Since the time constant ΔT is determined by dividing the inductance by the total loop resistance, the following approximate relationship exists:

$$D = 100B\ L/R = 100BLI/V$$

where R = total loop resistance (ohms)
L = inductance (henries)
I = closed-loop current (amperes)
V= open-loop voltage (volts)

A typical selector magnet (L = 700 mH) will generate (a) less than 2 percent of spacing bias when a 120 Vdc battery is used in a 60-mA, 45-baud loop, and

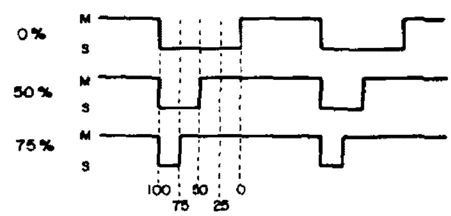

fig. 2. Degrees of marking bias.

(b) nearly 15 percent of spacing bias when operated at 13.2 volts (the minimum amount capable of supplying the 60-mA current through a 110-ohm coil).

Keying relays. When properly adjusted, relays can be useful in controlling dc loop circuits. The inability to follow fast keying pulses instantaneously causes relays to introduce significant delay to RTTY pulses – but this delay need not generate bias distortion if applied equally to making and breaking events.

Polar relays meet this requirement, but their use should be limited to cases where adequate instrumentation is available to keep the relays adjusted properly. Reed and mercury-wetted relays are sufficiently stable to requre no adjustments, but their selection must be accompanied by care in design since the make and break times of these relays almost always differ. **Fig. 4** shows one way to compensate these devices. R1, C1 are selected to protect the contacts but may introduce some marking bias. CR1 protects the transistor from over-voltage when the relay coil is switched off. Increasing R2 will decrease marking bias, while increasing R3 will decrease spacing bias. This

latter operation may require a higher A+ voltage or lower coil voltage.

FSK keyers. Bypass capacitors are a requirement in the design of FSK keyers, but care must be taken to ensure that the dc time constants of the keying circuit

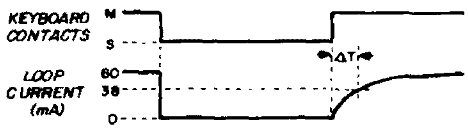

fig. 3. Spacing bias due to low loop battery voltage.

are small compared with the signaling unit time (1/B). An FSK keyer that works *fine* at 60 *wpm* may present excessive bias distortion at 100 wpm.

Transmitting contacts. Some transmitting contacts, such as the Model 14TD and Model 32, have accurate, distortion-free distributors when properly operated. Model 28 keyboards and TDs require careful setting of a single adjustment screw to eliminate bias and normally generate only minute end distortion. Models 14, 15 and 19-series keyboards, however, can generate large amounts of bias and end distortion when any of the six contact springs are improperly adjusted, as shown in fig. 5.

Speed error. The start-stop signaling system used in RTTY will tolerate nominal speed error. One way to look at speed differential is as if it were distortion that gradually increases as the signal progresses toward the end of each character. A page printer, for example, will react to a perfect signal sent too slowly, as if that signal had a distribution of spacing bias and marking end distortion.

FSK demodulators. One of the most severe causes of distortion can be the FSK demodulator. In some cases some of this distortion can be traced to poor design or adjustment in input, channel, or low-pass filters. The designer, for example, may have paid too little attention to phase response and the filter will have excessive overshoot and ringing. These situations generally will produce bias distortion, along with lesser amounts of end distortion, that are due to time constants and settling times in excess of one unit time (1/B). Such distortion often varies widely, depending on the characters received.

receiving considerations

The most limiting form of received distortion is generated by noise, interference, and signal fading. Each dB of improvement in this area calls for increased complexity and refinement. However, an occasional error will occur even with the strongest signals. Since noise can occur at any point on the signal waveform, a high probability of all types of distortion can be expected. To perform best with this random type of distortion, the receiving teleprinter should sample each signaling unit as close as possible to

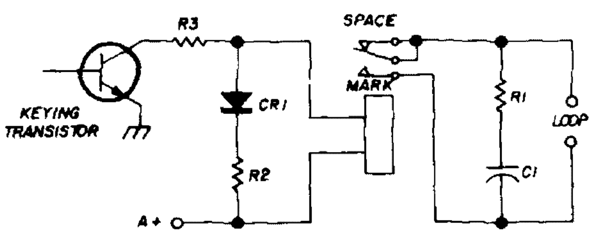

fig. 4. One way to compensate a nonpolar keying relay. Similar techniques can be used to compensate a transistor-keyed selector magnet.

its center. In practice, each receiving RTTY device must be adjusted for a maximum, but balanced, response to each of the four pure distortion types.

distortion test equipment

Common test equipment can be used for a few distortion measurements when specialized equipment isn't available. A Model 15 keyboard may be adjusted, for example, with an overdamped vom as follows.

Connect the vom across the keyboard in the low-ohms position to minimize the effect of the filter network. Adjust the ohms potentiometer for full-scale. Ob-

serve the 0 to 10-volt scale. Check that the meter reads 0 volts when the keyer contacts are open. Now turn on the motor and cause the clutch to engage steadily for repeated sending. Encode the BLANK symbol and adjust the stop spring for a reading of 1.92 volts. Now encode the characters E, LF, SPACE, CR and T, one at a time, while adjusting springs 1 through 5 for 3.26 volts.

The Model 28 signal generators may be adjusted similarly by repeat sending of the character R and adjusting for a reading of 4.62 volts.

A general-purpose oscilloscope with a triggered sweep may be used to analyze pure bias distortion; thus, it can be useful in adjusting relays, Model 28 contacts, and for resolving many other problems. A general-purpose scope is of little value, however, in the measurement of end distortion or bias in the presence of end distortion. This is because an ordinary scope won't synchronize on only the start pulse unless the sweep rate is adjusted for

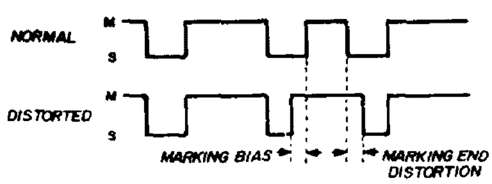

fig. 5. Bias and end distortion on the letter J from a Model 15 keyboard with a misadjusted number 4 contact spring.

one sweep per character, in which case the resolution is not sufficient to read distortion with meaningful accuracy.

special-purpose test gear

One of the best types of distortion test sets for RTTY received-signal analysis is the Western Electric 164C. This instrument has a special-purpose scope display calibrated linearly from 0 to 50 percent. The display is idle when no signal is present, and, after encountering a start pulse, a waveform generator causes seven horizontal linear transistions across the screen at the same rate in both directions (fig. 6).

The vertical deflection plates are fed from the input signal through an RC network that causes a mark-going bias edge to appear as a pip above the scope centerline and a space-going end-distortion edge to appear as a pip below the line. Marking and spacing distortion

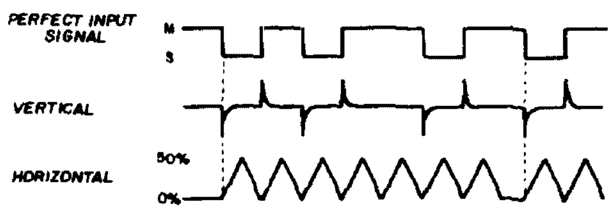

fig. 6. Typical waveforms for received distortion test sets of the type similar to the Western Electric 164C.

may be determined by the sweep direction at the time of the distorted edge, as shown by the shape of the pip. Typical sample signals are shown in fig. 7.

This type of instrument is ideal for setting up transmitting contacts, polar relays, and for optimizing demodulator design. Also, since it reads directly on any received message at either tape or keyboard speed, it may be used to analyze received RTTY signals during normal station operation.

test-message generator

Another useful instrument is the test-message generator. The best of these will generate a complex message consisting of all possible characters, with switched capability for accurate generation of the four types of distortion. Some units, like the Teletype DXD-100, combine the ability to generate a fox message with continuously variable distortion to 100 percent, with a stroboscopic display that may be used to adjust transmitting contacts. This instrument isn't very useful for received signal analysis, however, unless the incoming message is at tape speed and exactly at the same speed as the test set.

These distortion-generating test sets allow optimization of receiving teleprinter equipment. The best procedure is

to print the test message while gradually decreasing distortion until perfect print is obtained for one or two lines. This procedure is repeated four times, once for each distortion type. The amount of tolerance to each distortion type should be logged, along with the position of the range finder on the teleprinter under test.

The procedure described is repeated for several different range positions until the best balance of distortion tolerance is obtained. That range setting is then locked into the machine and the worst-case distortion tolerance evaluated. Machines not able to tolerate, say, at least 35 percent of any type of distortion without errors at this final range setting may need adjustment or lubrication. Model 28 machines typically tolerate over 40 percent of any type distortion.

Many other types of distortion test sets are available. Some have peak-reading meter readouts; others have digital distortion displays. Each can be a useful addition to the RTTY station.

regenerative repeaters

The regenerative repeater, once a troublesome vacuum-tube device, can now be built easily with only a few integrated circuits. Regenerative repeaters receive the incoming signal from the demodulator and regenerate a perfect signal, which is fed to the teleprinter. Since the highspeed ICs can sample 50 percent of the time with great accuracy, generally no range adjustment is necessary. These circuits can often be used to increase the distortion tolerance of the printer, while easing the requirement for accurate range adjustment on the teleprinter.

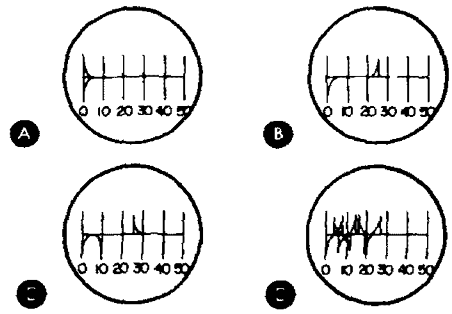

fig. 7. Perfect signal (A). 25-percent pure marking bias (B). 25-percent spacing bias plus 10-percent spacing end distortion (C). Demodulated signal received in the presence of noise, including random distortion peaking at 25 percent (D).

Probably the most essential requirement of a regenerative repeater is an accurate time base, stable with temperature (preferably crystal controlled), and capable of jitter-free start-stop control. Where these conditions can be met, the regenerative repeater is well worth installing as part of the FSK demodulator.

conclusion

The need for measuring and eliminating the various forms of pulse distortion have been around for as long as people have been communicating with Morse code. Telephone companies have elaborate test stations to maintain landline telegraph systems. Many of the procedures developed over the years can help make a science out of a situation that may have meant nothing but unexplained and irritating errors to many RTTY amateurs.

ham radio

"And two large barrels, a coil of copper tubing, kerosene burner, two dozen jugs. . ."

advanced divide-by-ten frequency scaler

This simple 10:1 prescaler will increase the frequency range of your counter to 300 MHz

F. Everett Emerson, W6PBC, 1709 Notre Dame Avenue, Belmont, California 94002

A **ten-fold increase** in the range of many frequency counters is possible with the use of the simple 10:1 prescaler of advanced design described here.

Frequency measuring systems, some simple, some highly sophisticated, have stirred the interest of hams for many years, but never more enthusiastically than in recent years when digital frequency counters have become available. The usefulness of such systems in the ham shack and workshop has made them exceptionally popular. Indeed, their popularity has become so widespread that frequency counters are even available in kit form.

Since becoming intrigued by frequency-measuring counters about two years ago, I searched through the ham literature for articles on counters and counter accessories.[1,2] In addition, two Heathkit counters were closely observed and their frequency limits and sensitivity were carefully noted.* K4EEU's divide-by-ten scaler, using four ICs, was constructed; when used ahead of the Heathkit counters, it permitted measurements up to 106-MHz. This appeared a worthwhile combination. In fact, it was a very happy combination which sufficed until Heathkit came out with its preassembled 80-MHz counters and a 175-MHz scaler kit.[3]

Then, dissatisfaction with the 106-MHz setup quickly took hold. As an end result, a 9-digit counter was built which had an upper frequency limit that turned out to be 125-MHz. A search was then made for a better scaler to further extend this upper limit.

My attention was drawn to Fairchild's 9500 series high speed, emitter-coupled

*While Heathkit advertises a frequency limit of "over 15 MHz," both counters measured well above 15 MHz, one reaching 25 MHz, and one reaching 30 MHz.

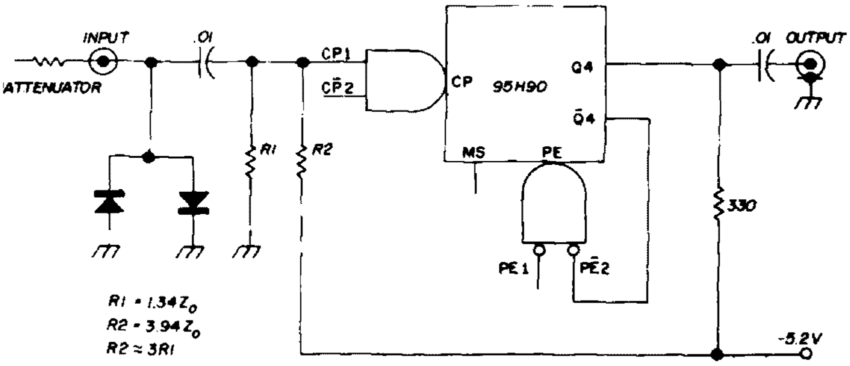

fig. 1. Logic symbol and connections to the Fairchild 95H90 IC used in the 300-MHz prescaler.

logic (ECL) integrated circuits. Among them is their 95H90, a very high speed, *temperature* compensated, ECL circuit for frequency division. This is a divide-by-ten prescaler usable to about 300-MHz. For the owner of a counter whose upper frequency limit is on the order of 25 to 30 MHz, such a prescaler would permit measurements in every amateur band up to and including the 220-225 MHz band. The prescaler described here will do that and more. It should prove to be especially attractive to the vhf enthusiast.

circuit

Three prototype prescalers have been constructed using the Fairchild 95H90 for ten-to-one frequency division. As constructed (one hard-wired bread board and two one-sided foil circuit boards), their lower limits for sine-wave inputs were found to be between 6 and 9 MHz and their upper limits were between 220 and 272 MHz. These variations are apparently due to slightly different construction techniques, types of component capacitors used and possible variations in the ICs. All, however, were decidedly successful and exceedingly stable.

Since the output is a square wave, the output will be accepted by practically any counter. In this connection, it should be noted that the output is deliberately made through a dc-blocking capacitor, a *must* if the divice is to be used ahead of the Heathkit counters whose inputs do not have one.

The Fairchild 95H90 is a high speed ECL MSI device, designed specifically for the communication and instrumentation industries. All of the high-speed logic manipulations are "on chip." In this single, 16-pin, dual-in-line IC package, frequency division by ten may be quite simply accomplished and at a lower over-

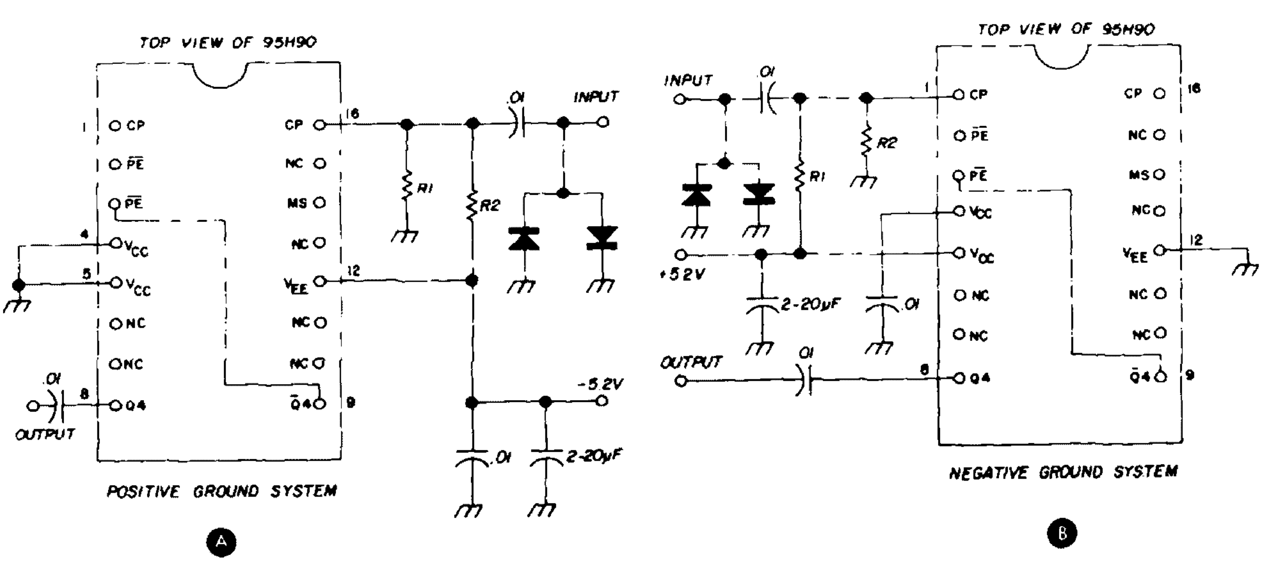

fig. 2. Complete schematic of the 10:1 prescaler. Circuit in (A) is for positive ground. Circuit in (B) is for negative ground. *For 50-ohm input,* R1 = 68 ohms and R2 ≈ 200 ohms.

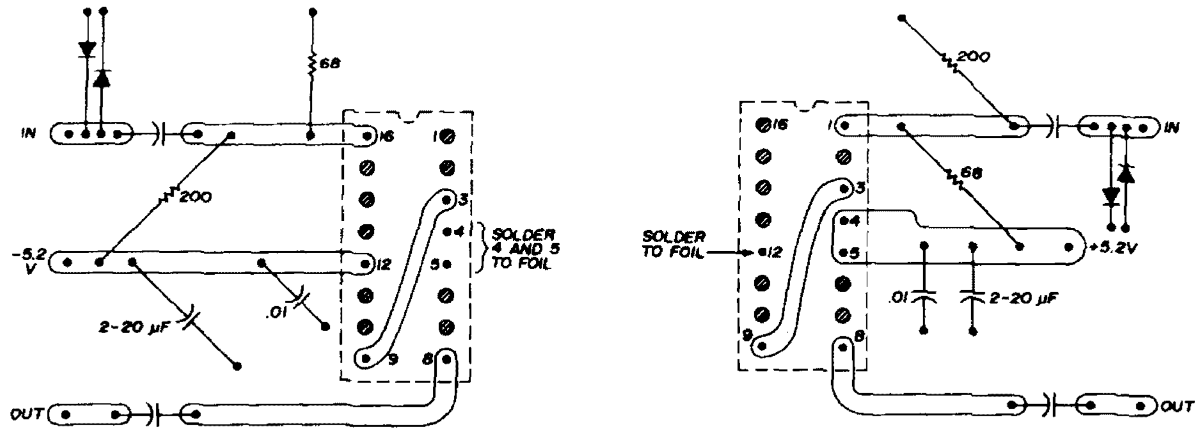

fig. 3. Circuit-board layout for the presealer is extremely simple. Layout in (A) is for positive ground; (B) is for negative ground.

all cost than scalers using a multiplicity of ICs which divide by two, and then, by five. Although the 95H90 may be so connected to divide by eleven, and with other control logic a divide-by-N counter can be constructed, these uses are not of concern here; only the divide-by-ten connection will be considered.

The 95H90 prescaler logic symbol and its necessary external components are shown in fig. I. A well-filtered and regulated 5-volt power supply is required. Resistors R1 and R2 set the input bias and are in the ratio of R2 equaling approximately three times R1. These resistors also partially determine the IC's input impedance, and are specified as R1= 1.34Z, and R2= 3.94Z. Thus, for a 50-ohn input, R1= 67 ohms and R2= 197 ohms. For practical usage, R1 may be 68 ohms and R2 may be 200 ohms for an input impedance of 51 ohms.

The schematic of the 10:1 prescaler is shown in fig. 2. Fig. 2A is for use where the positive supply voltage is connected to the circuit board foil (positive ground), while fig. 2B shows the connections where the supply negative is connected to the board foil (negative ground). Fairchild specifies the positive ground; however, each connection (positive or negative) has been tried on separate prototype boards with no discernable ill effects on the device's frequency limits. The negative ground seems to be preferred by amateurs. Note that fig. 2 shows top views of the 95H90, in accordance with the practice of the IC industry.

construction

Circuit-board layout, showing bottom views (foil side), are shown in fig. 3 The simplicity of these layouts should be readily apparent. For those who can make them, etched boards are ideal. Because of the circuit simplicity, however, the prototype boards were made by first drawing enclosing lines on the boards with a marking pen and then gouging away the lines with a hand-held

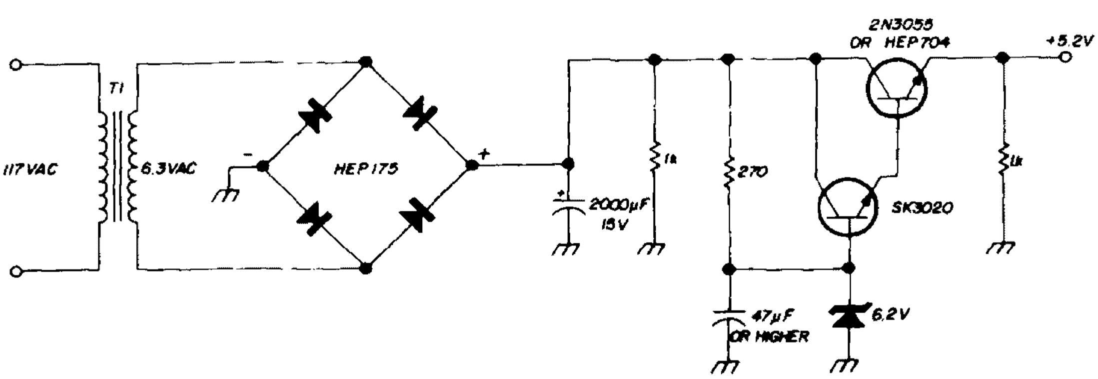

fig. 4. Power supply for the prescaler provides 5.2 Vdc output.

hobby-type burr drill, a process which took about five minutes. For a neater job, lines may be readily removed on a drill press.

In any event, the prototype method proved so acceptable that one of the original boards is now in permanent use. Fig. 3A shows the positive ground, while fig. 3B shows the negative ground.

Half-watt resistors were used because of their ready availability. Quarter-watt resistors would permit an even smaller layout. Capacitors should be of the best quality. The .01-μF capacitors should be ceramic discs because these types offer lower impedances than other types. The large bypass capacitor (2 to 20 μF) should be a tantalum type. If a tantalum capacitor is not available, use a physically small electrolytic.

In one prototype a 10-μF, foreign-make, electrolytic was used, along with Mylar-type .01-μF capacitors, without deterioration of more than a few MHz in the upper frequency limit.

Note the protective diodes at the prescaler input; use the fastest diodes you can obtain. They are a cheap way to save a moderately-expensive 95H90.*

A suitable power supply is shown in fig. 4. Ready availability of parts determined its capacity, not the need for large currents. The prescaler draws less than 200 mA. Fairchild's 95H90 data sheet specifies a 5.2-volt supply. The power supply voltage tolerance should be held within plus or minus 5% of 5 volts.

One prototype prescaler optimized at 5.2 volts and another at 4.8 volts; however, the standard supply is 5.2 volts and will service numerous ICs. Remember that the output voltage will equal the zener voltage minus the drop in the two transistors. Thus, with a 6.2-volt zener, the output will be 5.2 volts.

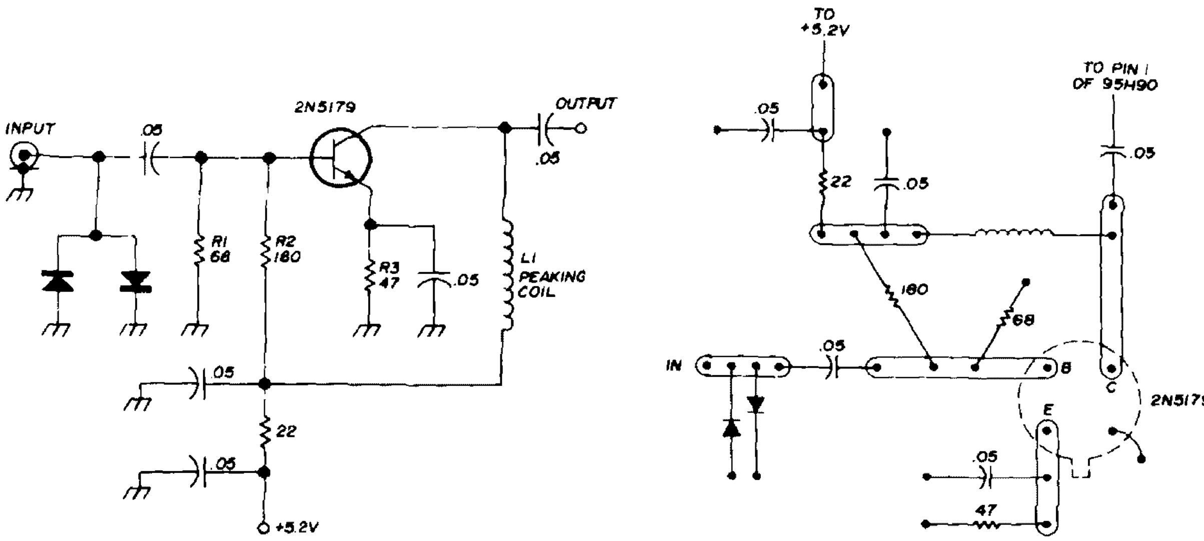

fig. 5. Simple preamplifier circuit increases sensitivity of prescaler. Peaking coil L1 is 8 turns no. 26, 5/32" diameter, air-wound, spread to cover 5/8" length. Circuit-board layout is shown to right.

sensitivity

The sensitivity of the prescaler prototypes varied from approximately 130 millivolts at 100 MHz to 240 millivolts at 260 MHz. This is adequate for many uses. For greater sensitivity, however, a single transistor preamplifier may be added. A schematic of a wideband amplifier, using a 2N5179 transistor, which is suitable for use with this prescaler, is shown in fig. 5. Normal vhf construction practices, such as the shortest possible leads and ade-

*When ordering the 95H90 IC, specify the Fairchild U6B95H9059x. Circuit Specialists, Box 3047, Scottsdale, Arizona 85257 has all the semiconductors for this unit in stock. The 95H90 is $16.00, the 2N5179 is 50 cents, the HEP 704 (2N3055) is $2.50 and the HEP 175 is $1.35. Please add 35 cents for shipping.

quate decoupling, are needed.

If this preamplifier is used with the prescaler, R1 in fig. 2B should be changed to 4700 ohms, and R2 should be changed to 1500 ohms. Preamplifier sensitivity will range from about 15 millivolts at 100

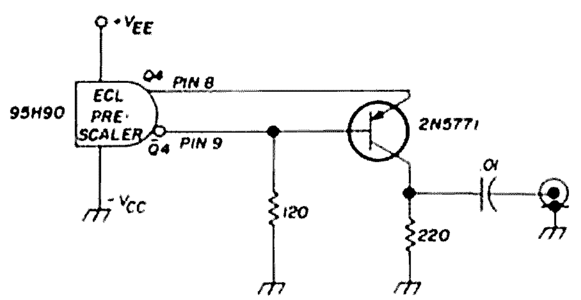

fig. 6. Post amplifier circuit for use only if counter has insufficient sensitivity or if a TTL interface is required.

MHz to about 100 millivolts at 220 MHz. All in all, a worthwhile addition.

If your present counter is deficient in sensitivity, a simple single transistor post-amplifier may be added to the prescaler. Its schematic, using a 2N5771, is shown in fig. 6 for a negative-ground supply. However, a counter which would not respond to this prescaler is yet to be found. If, by chance, yours does not, it's high time to find out what is wrong with your counter. The amplifier shown in fig. 6 makes an excellent interface if such is desired for connection to a following TTL device.

This article could not be considered complete without expressing sincere appreciation for the valuable assistance and checking of the devices by Robert Melvin, W6VSV. I am deeply grateful for his help and his enthusiastic support.

references

1. Bert Kelley, K4EEU, "Divide-by-Ten Frequency Scaler," *ham radio,* August, 1970, page 26.
2. Kenneth Macleish, W7TX, "A Frequency Counter for the Amateur Station," *QST,* October, 1970, page 15.
3. Factory-built Heathkit 80-MHz counters include the model SM-104A ($540) and the model SM-105A ($375). The Heathkit 175-MHz scaler kit, model IB-102, sells for $112.95.

ham radio

repeater control with simple timers

All-electronic timers provide reliable, simple and inexpensive repeater control

George R. Allen, W2FPP, and Richard J. Sobus, K2QLE

Repeater activity is increasing daily, and it seems that there is always need for one more repeater in town. Unfortunately, installing a repeater usually turns out to be quite a difficult project, and for every operational repeater on the air there are probably ten more that will never make it. The unfinished repeaters usually run into many roadblocks, one of which is the problem of designing and installing the necessary controls such as delay timers, three minute timers and repeater ID timers.

From our experience, it appears that a lot of the repeaters today are using mechanical timers with their inherent disadvantages of high cost, inadequate reliability, large size and susceptibility to adverse environmental conditions. The timers described in this article were designed to provide a means for eliminating the roadblocks caused by using the con-

Printed circuit boards for the repeater control unit are available from Alton Industries, 7471 Thunderbird Road, Liverpool, New York 13088. *Drilled boards are $4.50, undrilled* boards are $3.50. Included with the boards is a detailed schematic and board layout. Wired and tested units are available from the same source for $33.50.

ventional approaches to timers. These timers are highly reliable, simple to build, very inexpensive and can be connected together to provide repeater control without need for mechanical contacts. This article also shows the complete interconnection of these timers to form a repeater control unit.

The two timers shown are based on the same principles; however, there is one basic difference. The timer in **fig. 1** will deliver one output pulse *after* a given time interval, while the timer in **fig. 2** will deliver the output pulse *before* the time interval. In both cases, only one output pulse can occur for each time interval regardless of the number of input pulses. In both cases, the timing cycle will not repeat unless an input pulse is received.

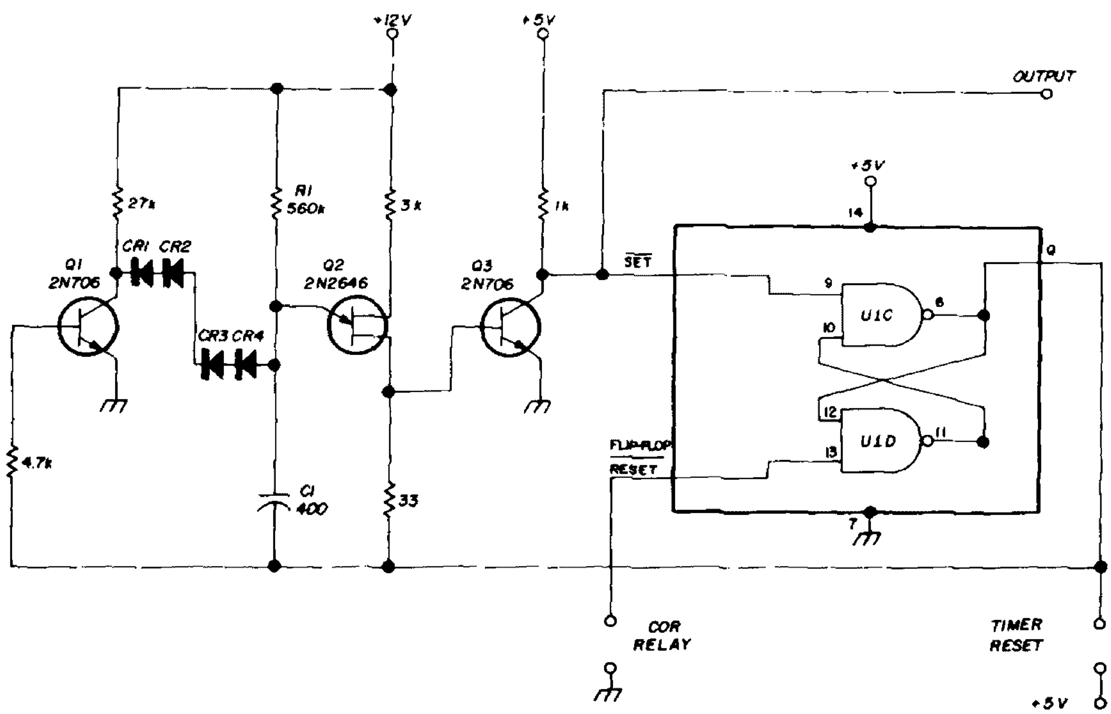

fig. 1. Basic timer to give one pulse after a given time interval. U1 is a 7400 IC. CR1 through CR4 are silicon diodes. R1 and C1 can be changed to provide different timing intervals.

timer operation

When power is first applied, C1 begins to charge through R1 until the emitter of Q2 reaches approximately 6 V. At that time the 2N2646 unijunction fires, resulting in a positive pulse across the 33-ohm resistor. This pulse is inverted to a negative going pulse by Q3. With the constants shown, the approximate timing cycle is two minutes and ten seconds. A *2N2647 is a more desirable unijunction*, but the 2N2646s were on hand and gave satisfactory performance.

In the circuit of **fig. 1**, the output pulse triggers a set-reset flip-flop which turns the timer off. The timer will be started again by grounding the reset line of the S-R flip-flop. The *timer reset* is used to set the timer to its initial state. It will do this even if the carrier-operated relay is closed. Its main function is to reset the timer if it is not desired to complete the timing cycle.

In the circuit of **fig. 2**, the output pulse sets the S-R flip-flop which then results in a +5 V level at pin 5 of the NAND gate U1B. In order to get an output from this timer, a zero volt output is required at pin 6 — this output can be obtained only when both inputs are at a

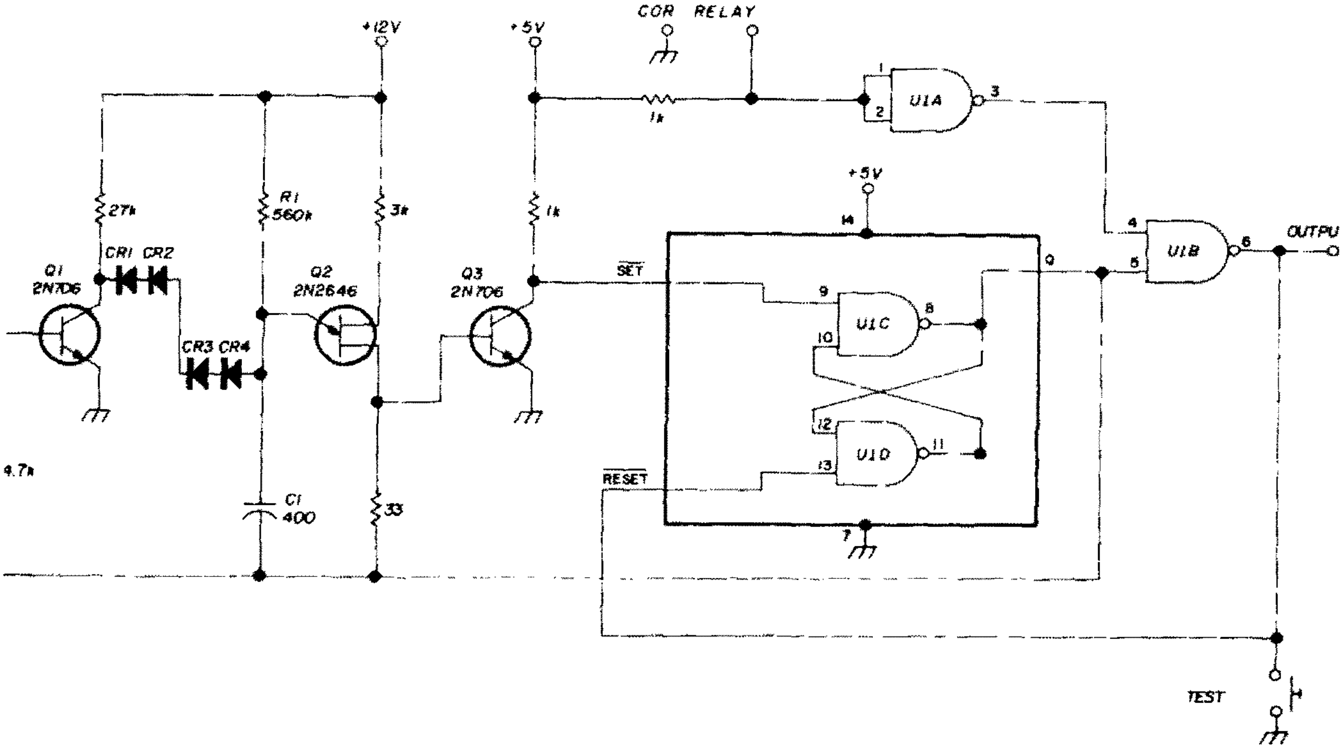

fig. 2. Basic timer to give one pulse before a given time interval. Other notes are the same as in fig. 1.

+5 V level. As connected, U1A is an inverter which has a zero volt output when the input is open and a +5 V output when the input is grounded. Therefore, when the input is grounded, both inputs to U1B are +5 V, the output goes to zero and the flip-flop resets. The resultant output is a pulse at the beginning of the timing interval. With this timer, an output can occur only when the COR contacts are closed. Thus, if this timer is used for an identifier, the ID will trigger only when the repeater is active.

timer applications

The timer in **fig. 1** can be used as the

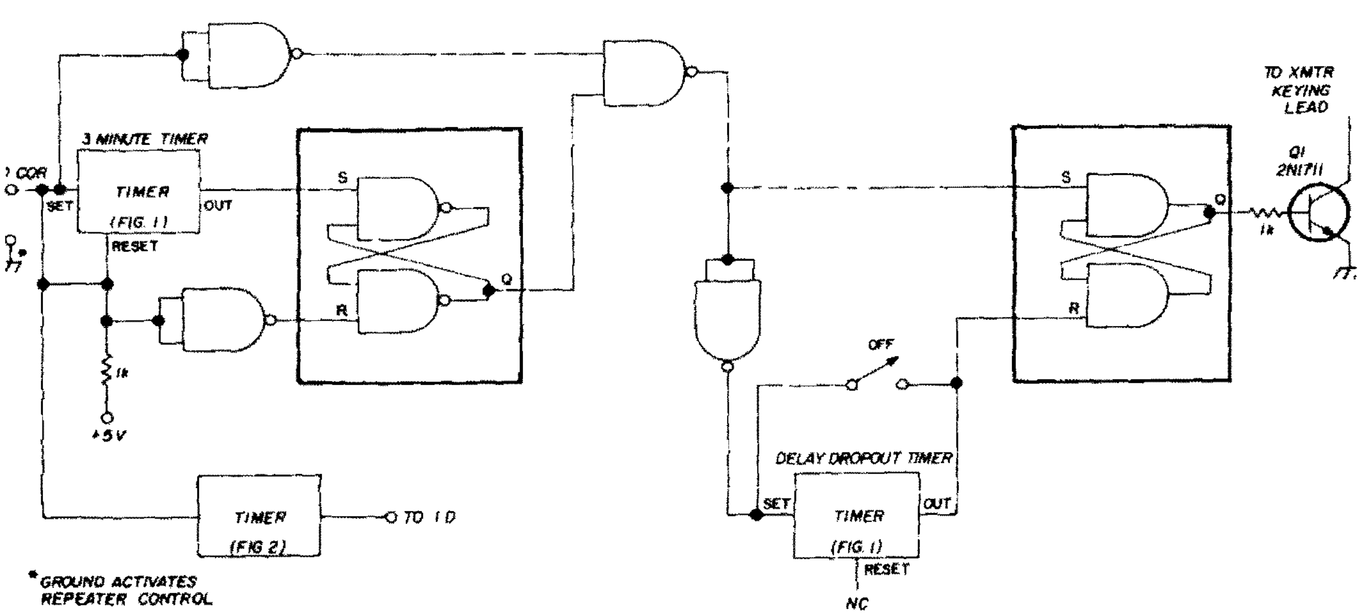

fig. 3. The complete repeater control unit. The gates are all parts of 7400 integrated circuits.

three-minute timer for a repeater by selection of the proper value of C1. The input can be the contact closing of a carrier-operated relay, while the output pulse can trigger a simple S-R flip-flop which will disable the transmitter. Thus, if the COR is on for longer than the specified interval, the transmitter will be timed out. This timer can also be used to provide a drop-out delay for the transmitter. In this case, the absence of carrier will trigger the timer and the resultant delayed pulse will cause the carrier to drop out.

The timer in fig. 2 can be used to trigger a repeater identifier. In this mode of operation, the ID will only go on at the beginning of a timing cycle. If the repeater has been inactive for longer than the timing interval, the ID will be keyed up when the COR is keyed, and the ID will be keyed every two minutes and ten seconds as long as the COR is keyed. The ID cannot be keyed unless the COR is keyed, thus eliminating the problem of the ID keying up the repeater when there is no activity.

repeater control unit

The interconnection of the timers to provide a complete repeater control with a three minute timer, a drop out delay timer and a repeater identifier timer is shown in fig. 3. The entire circuit as shown was used at the WA2ZVZ repeater until it was replaced recently by a commercial unit with telephone-line remote-control capabilities. The ID timer portion, however, is still in use. There have been no failures of this timer with temperatures ranging from sub-zero to 80 degrees. The only phenomenon noticed was a decrease in the timing cycle amounting to about five or ten percent during extremely cold weather.

The circuits presented in this article are reliable, simple to build and inexpensive. it is hoped that these circuits will help many repeaters become operational by overcoming some of the problems relating to timers and repeater control.

ham radio

solid-state hang agc circuit for ssb and cw

A high performance solid-state hang agc circuit for use with fet and mosfet rf circuits

Jerome L. Hartke, W1ERJ, 119 Fairbank Road, Sudbury, Massachusetts 01776

In the rather distant past, when amateur radio operators depended upon a-m for communications, automatic gain control in the receiver was a simple matter. The negative dc voltage necessary for agc was easily derived by sampling the output of the diode a-m demodulator. This negative voltage was fed to the grids of the rf and i-f amplifier tubes to maintain a constant detected audio level which varied only slightly as the signal faded or as various stations were tuned in, and was also used to control the S-meter. Due to the presence of a carrier, agc circuits for a-m were relatively simple. Subtle factors such as attack and decay time could be comfortably ignored. The CW gang, purists that they were, happily rode the manual rf gain control while they pounded brass.

Single sideband brought an end to these happy days, and in the few cases where the local bfo signal did not pin the S-meter, the absence of a carrier during periods of no modulation caused S-meters to flop erratically. With the introduction of product detectors, which kept the bfo signal away from the agc detector, gain-control circuits were used which permitted fast rise and slow decay of the dc output. However, to hold the agc level and receiver gain reasonably constant between voice syllables or code characters, the decay-time constant had to be so long that receiver gain did not respond quickly enough to signal-level changes experienced with fading signals or roundtables.

A nearly ideal ssb and CW agc circuit

was introduced by W1DX in 1957.[1] This circuit separated the agc line from the "clock" which controlled the hold time. This development, dubbed *hang agc*, allows the agc line to remain at a steady negative voltage following a voice syllable or code character; the "clock" counts off a predetermined time, after which it triggers a discharge circuit which rapidly restores the receiver gain to maximum in the absence of a signal.

The W1DX circuit requires three diodes, one triode and a voltage step-up transformer along with some resistors and capacitors, and its power and space requirements are substantial. Nonetheless, it has been used to supply audio-derived

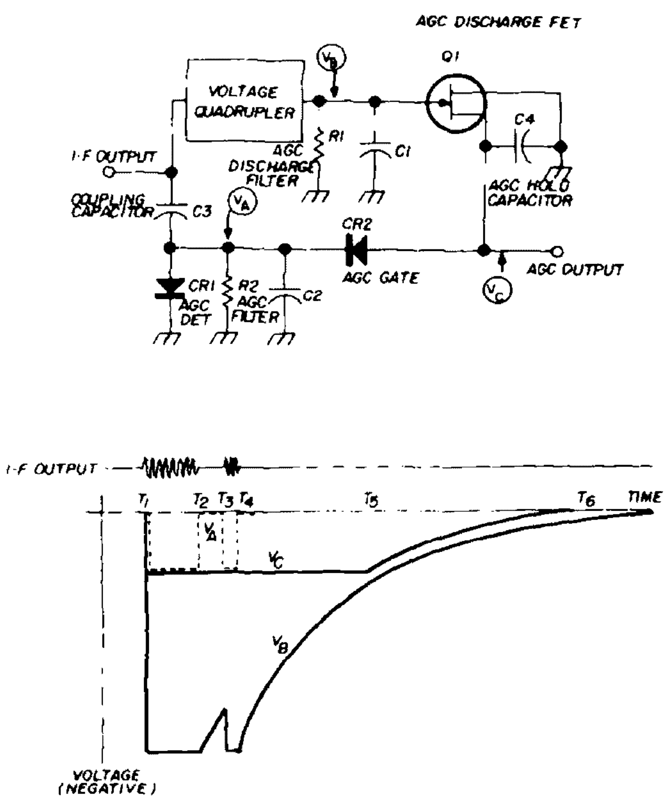

fig. 1. Basic solid-state hang agc circuit and dc voltage levels at different times in the operating cycle.

performance has easily justified the extra parts. Since a high-impedance agc line is necessary for the hang circuit, it does not interface well with bipolar transistorized receivers unless a buffer amplifier is used.

The recent popularity of field-effect rf

table 1. Hang and discharge times for the circuit of fig. 1. Times are in seconds for resistance in megohms and capacitance in microfarads. Vp is the pinch-off voltage of Q1.

$\frac{V_A}{V_P}$	hang time (T5-T4)	discharge time (T6-T5)
0.5	0.29 R1C1	0.41 R1C1
1.0	0.69 R1C1	0.69 R1C1
2.0	0.98 R1C1	1.1 R1C1
4.0	1.16 R1C1	1.61 R1C1
10.0	1.29 R1C1	2.40 R1C1
20.0	1.34 R1C1	3.05 R1C1
40.0	1.36 R1C1	3.71 R1C1

and i-f amplifiers, particularly those using dual gate mosfets, has revived the high-impedance agc line. The circuit described here is an all solid-state version of W1DX's hang agc system which is small, inexpensive, requires no external power and can be used with either tube or fet rf amplifiers. It operates directly from the i-f strip, avoiding the minor difficulties which occur in many audio-derived agc systems.

basic circuit

Fig. 1 shows the basic circuit and the dc voltages at various stages of its operating cycle. When a signal appears at the i-f output at time T1, it is rectified by CR1, filtered by R2C2, and gated as a negative voltage onto the agc hold capacitor C4, and the agc line. Simultaneously, a negative voltage four times greater than the agc level is developed across C1 which prevents Q1 from conducting. When the signal is removed at time T2, voltage V_A rapidly drops to zero. However, the agc line voltage, V_C, does not drop because Q1 is off, CR2 is reverse biased and the agc line has no dc return to ground (resistance to ground should be greater than 100 megohms for the circuit to work properly). Capacitor C1 begins to discharge through resistor R1, causing the gate voltage of Q1 to drop. Before Q1 can conduct, however, the signal reappears at time T3, recharging C1, C2 and C4.

Following the disappearance of the signal at time T4, V_A drops to zero and V_C, the agc level, remains constant while V_B decays with the *time* constant R1C1. At time T5, the negative gate-source

voltage of Q1, V_C-V_B, has reached the pinch-off voltage of the fet, V_P or $V_{GS(off)}$, and Q1 conducts, discharging the agc hold capacitor. The agc line goes to zero volts in the time interval T5 to T6.

Reviewing the operation of the circuit, CR1, R2 and C2 maintain a negative dc voltage, V_A, which instantaneously follows the envelope of the i-f signal. The agc gate, CR2, allows the hold capacitor, C4, to charge up to the peak value developed across C2 and to hang at that value after the signal decreases in amplitude. The voltage quadrupler maintains a voltage V_B, approximately four times V_A, which biases Q1 into cutoff. Voltage V_B rises very fast when a signal appears, but decays quite slowly after the signal disappears. This slow decay, with time constant R1C1, is the "clock" that controls the hang time (the interval T5-T4 in fig. 1). After the prescribed amount of "clock" time has elapsed, the agc line is discharged if there is no signal, and the receiver returns to maximum gain.

By properly choosing the RC time constants in the circuit, receiver gain is quickly reduced as a signal appears (no unpleasant thumps caused by slow attack time constants). The gain then stays absolutely constant between voice syllables or code characters (the S-meter doesn't even wiggle), but recovers rapidly enough after the hang time to follow fading signals or weak stations in a roundtable.

The circuit has an intrinsic threshold, below which weak signals and noise will not activate the agc. The threshold level is approximately one volt, and is equal to the forward drop of CR1 (0.7 volt) plus $V_P/4$, (V_P is the pinch-off voltage of Q1). Although the hang and discharge times of the circuit depend slightly on signal level, the variations are scarcely noticed during operation. Exact values are given in table 1.

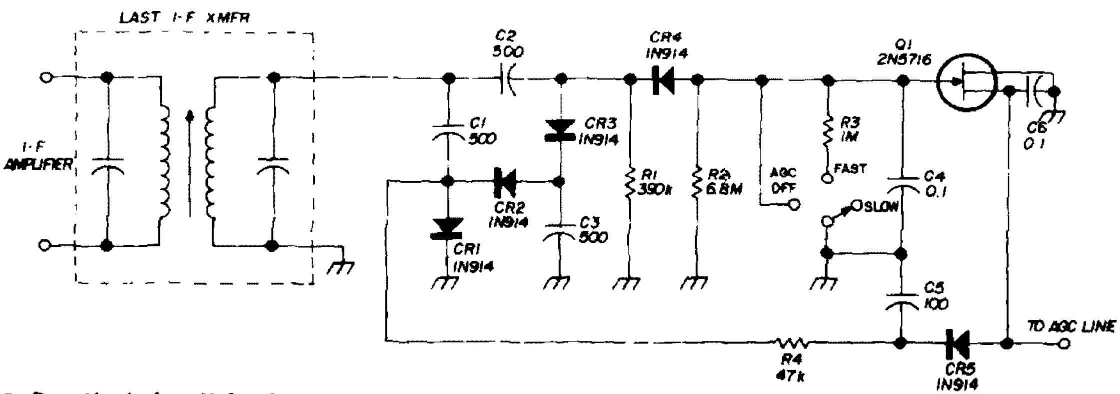

fig. 2. Practical circuit for hang agc.

practical circuit

A working hang agc circuit with a total parts cost less than $4.00 is shown in fig. 2. Diode CR1 is the agc detector whose attack time is controlled by C1 and the source impedance of the last i-f transformer. For reasonable source impedances of 10k or less, the attack time is shorter than 10 μsec. The decay time of the detected signal is also very fast, set by C1, C5 and R1 to 250 μsec.

The detected signal is gated onto the agc hold capacitor, C6, through R4 and CR5 with an attack time, R4C6, of 5

table 2. Dc voltages for the circuit of fig. 3.

test point	dc voltage
V1	-30 Vdc
V2	+4.3 Vdc
V3	+3.3 Vdc
V4	+12 Vdc
V5	+1.0 Vdc
V6	+30 Vdc

msec. This keeps C6 from being charged up by large, short-duration noise spikes. A negative voltage is developed by the voltage quadrupler, CR1-CR4, C1, C2, C3 and R1; this negative voltage is applied to the agc discharge fet, Q1. Resistor R1 causes the dc voltages across C1, C2 and C3 to decay with a 200-μsec time con-

CR4 and CR5 are particularly critical in regard to high back resistance. Most silicon computer diodes will work well, but conventional silicon detector diodes, silicon rectifiers or germanium diodes should not be used. The 1N914 types called for in fig. 2 are available at very reasonable prices.*

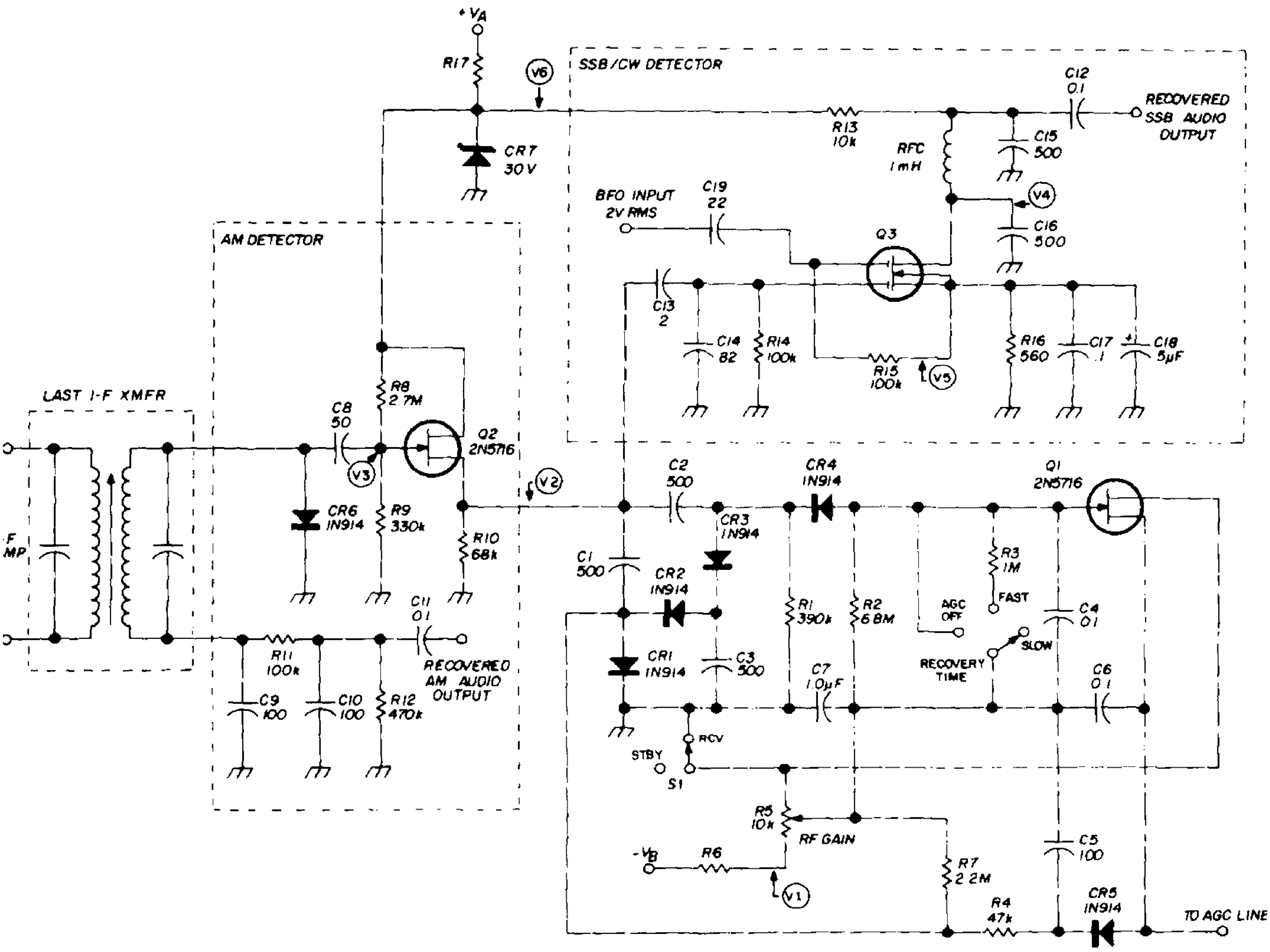

fig. 3. Hang agc circuit with a-m and ssb/CW detectors, manual gain control and disable feature for use when transmitting. Transistor Q3 is a Motorola MFE3008 or RCA 40604.

stant, allowing the quadrupler to recover quickly after removal of a signal. Hang time for ssb/CW operation is set at 0.7 second by R2 and C4, and can be reduced to 9 msec for a-m operation by placing R3 in the circuit. Agc is removed by shorting C4, which keeps Q1 in conduction at all times by grounding its gate.

All diodes in the circuit must be fast recovery types. In addition, they should have low reverse leakage currents. Diodes

The choice of an n-channel fet for Q1 is considerably narrowed by the need for a device having a low pinch-off voltage. Most commercially available devices have rather large pinch-off voltages in the 3- to 10-volt range. However, the Motorola 2N5716, which is available from

*1N914 diodes are priced at 16 for $1.00 from M. Weinschenker, Box 353, Irwin, Pennsylvania 15642.

Motorola distributors for $.80, has a pinch-off range of 0.2 to 3 volts. In addition, the 2N5716 is rated for up to -40 volts on either the gate or the source with the drain grounded. Many other fets are breakdown *rated* at about half of this value, prohibiting their use in tube-type receivers *where the* agc *voltage can* easily go -10 volts during normal operation, or where the output of the voltage quadrupler applies -40 volts to the gate of Q1.

expanded circuit

Fig. 3 shows the basic circuit of fig. 2 plus a number of features which add to the versatility of the hang agc circuit. Table 2 lists the dc voltages which should be observed at the identified points. The cold ends of R2, R3, C4, C5 and C6 have been lifted from ground and connected to the wiper arm of R5 which feeds a *negative bias to* the gate of Q1, and, through R7 and CR5, to the agc line. A drain current of less than one microampere, flowing through R7, keeps Q1 near cut-off. Resistor R6 should be selected to give -30 Vdc at point V1.

This manual gain-control circuit is most useful since, along with reducing the gain of the receiver, it provides an agc threshold.[2] If the S-meter reads the agc bias in the normal manner, then its reading will increase as bias is manually introduced, and the threshold prevents stations weaker than the level indicated on the S-meter from operating the agc circuit.

The entire agc line is put on standby during transmit with S1. Switch S1 removes the ground from R5 and, since R5 and R6 no longer act as a voltage divider, the full voltage, $-V_B$, is applied to the agc line. Since V_B may be in excess of 40 volts, the breakdown *rating of* Q1, its *drain is also* biased to $-V_B$ during standby to prevent damage to the fet. Switching action may also be achieved by connecting a set of transmit-receive relay contacts in parallel with S1.

Diode CR6 is the a-m detector and its output is filtered by C9, C10 and R11. The action of agc rectifiers CR1-CR4 would severely distort the detected a-m signal if they were connected directly to the i-f transformer secondary, thus the source follower Q2 is used to isolate the agc circuit.

A product detector, consisting of Q3 and its associated components, is used for ssb and CW. The dual-gate mosfet isolates *the bfo signal from the agc* detector, preventing agc action in the absence of any signal. The capacitive divider, C13 and C14, reduces the ssb/CW output to a level compatible with the output of the a-m detector.

Zener diode CR7 reduces the supply voltage, V_A, to a level suitable for Q2 and Q3. Resistor R17 should be chosen so that about 3-5 mA is drawn from point V_A. If a 30 ± 5 volt supply is already available, it may be connected directly to the junction of R8 and R13. Then CR7 and R17 are not required.

installation

Wiring of the circuit is non-critical except for the usual observance of reasonably short leads in the portions of the circuit which carry rf. Various methods of feeding agc voltage are illustrated in fig. 4. Existing receivers probably will require little or no modification to use these systems. The only precautions necessary are to keep the RC time constants small, being careful to use R and C values no greater than those shown, to preserve the rapid attack time of the agc circuit. In all cases the S-meter amplifier should have dc gate or grid characteristics similar to the devices used for rf and i-f amplification.

It is important to keep leakage resistance from the agc line to ground greater than 100 megohms. Smaller values will discharge the line more rapidly than intended. The input impedance of most voltmeters is too low to measure the agc voltage. The S-meter should be calibrated as a voltmeter if you want to measure the agc voltage.

The circuit shown will not operate with low-impedance bipolar-transistor agc lines. Buffer dc amplifiers must be used in such cases or in situations where the agc line must have a finite resistance to

ground. The design of such amplifiers is beyond the scope of this article, since each one must be uniquely related to the quiescent bias levels of the receiver it is used in.

The circuit of fig. 2 is presently in use in my NC-125 receiver, while the system of fig. 3 is functioning in my KWM-1. Reception of ssb or CW is excellent, with smooth agc attack and beautiful hang, followed by rapid recovery. It is impressive to watch an S-meter hold rock steady in the presence of a signal, yet quickly follow fading or the strengths of different signals while scanning the band. The hang agc system requires minimal space and power, and is inexpensive to build. Its use is highly recommended in new construction as well as in upgrading existing equipment.

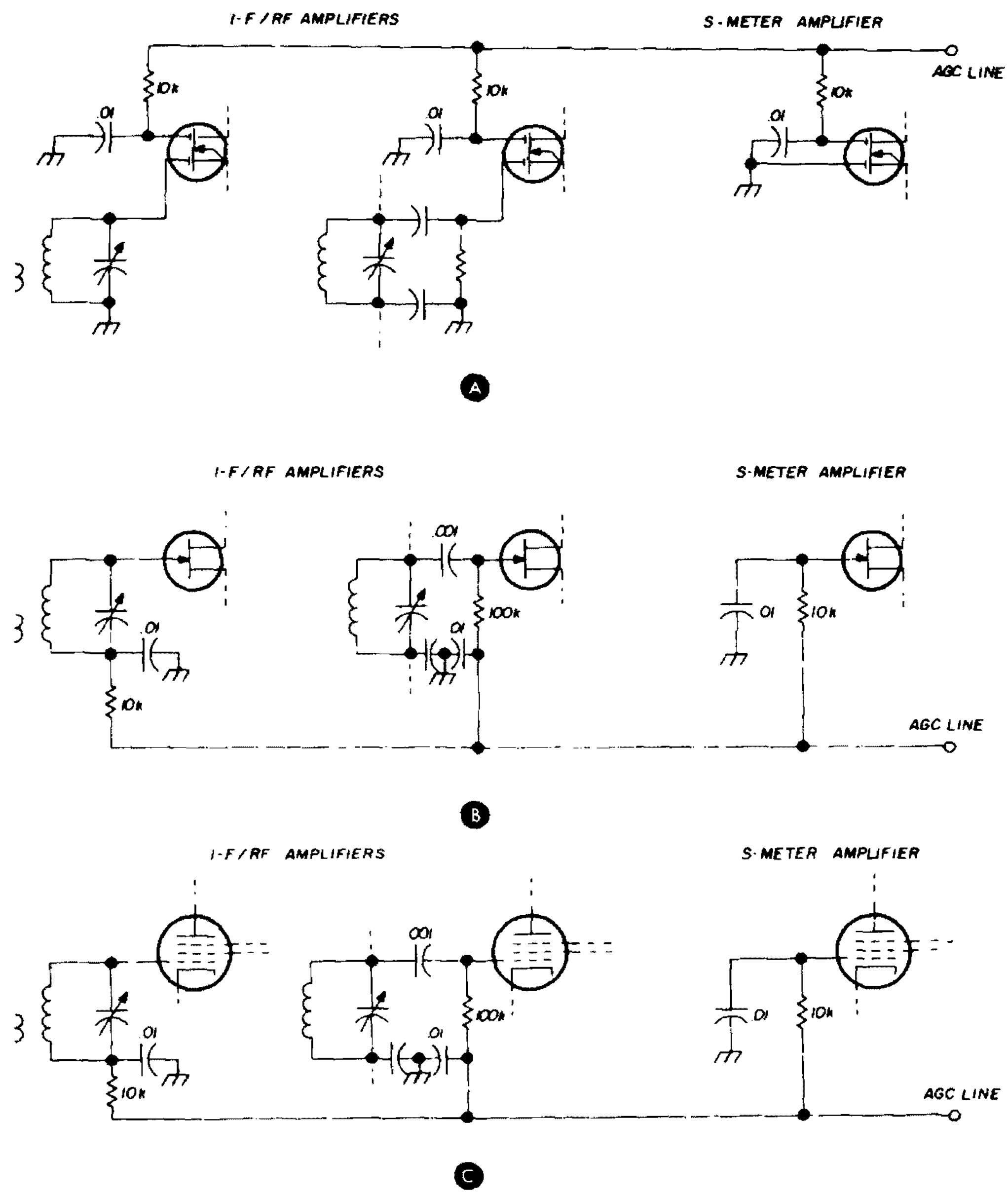

fig. 4. Typical agc distribution systems for dual-gate mosfets (A), jfets (B) and vacuum-tubes (C).

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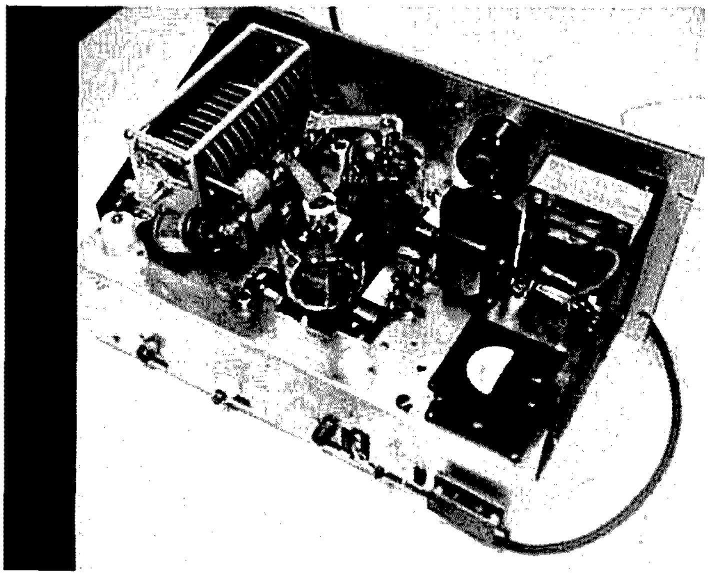

using odd-ball tubes in linear amplifier service

Practical suggestions for putting your surplus tubes to work

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If you have been an active amateur for any length of time, chances are you've accumulated several transmitting tubes. Such tubes may be triodes or tetrodes of various lineage — generally they have large envelopes, husky filaments, and are capable of withstanding quite high plate voltage. A challenging question is, how do you use these tubes in today's ssb linear amplifiers? Essential information is difficult to obtain for operating older tubes in linears.* This article offers some suggestions on using older transmitting tubes in a clean-sounding linear amplifier even if you don't have manufacturers' data on the operating characteristics of such tubes.

operating considerations

The specifications for modern triode linears indicate that only very high-mu tubes, preferably those that operate with zero bias, will be satisfactory. There is no doubt that such tubes are somewhat easier to put into service, but tubes with just about any range of amplification factor can be made to perform acceptably. In fact, at one time tubes with very low mu were considered best for linear amplifiers because they would take a wider swing of grid voltage before drawing grid current.

Present-day practice, which centers about the grounded-grid circuit, largely ignores such fine points. The very heavy negative feedback inherent in the grounded-grid configuration tends to iron out distortion products resulting from the

*The ARRL handbook lists ratings for most transmitting tubes. Some of the older editions contain data on special "one of a kind" tubes from WW II. editor.

sudden transition from no grid current to full grid current. Also, the fact that the rf excitation *fed through* the amplifier tube tends to keep a swamping load on the driver stage further lessens the transition effect. So don't be concerned if that oddball transmitting tube in your junk-box has low or medium mu.

You can do quite well without knowing the mu. All you really need to find out is what bias, at the plate voltage you intend to operate, is required to hold the plate dissipation to an acceptable amount. But remember that most tubes found on surplus will take a very impressive voltage and will approach linear operation more readily when the plate voltage is near the upper limit. You can determine this parameter if you have (or can borrow) a variable-voltage power supply that provides several hundred volts. In addition, you'll need a voltmeter of similar range.

determining operating parameters

Let's assume you've built your amplifier to the point where you're ready to start testing with the intent of arriving at operating parameters. Initially, you set the grid bias at the highest negative potential available, then you turn on the plate voltage. If all goes well no plate current will flow: so you gradually reduce the grid bias, keeping an eye on the tube plate. Plate current will start to flow as the grid bias moves out of the cutoff region. As plate current goes higher the tube plate will start showing color. Here is where you'll have to use judgment. If the tube has a carbon plate, a dull red at the plate center is the stopping point. If the tube has a tungsten plate, a bit deeper red around the center of the plate is permissible. If the tube has a tantalum plate, the red glow can be permitted to spread over most of the plate.

When the reduced grid bias (and the resultant plate current) has caused the plate color you've decided to accept, quickly turn off the plate voltage. Turn it on again for just a moment, and note the grid voltage. Again turn on the plate voltage just long enough to note the plate current. From the plate current and the plate voltage, compute the plate power input. If it's not over *one-half* the amount of the tube's rated plate dissipation, you're at a good starting point.

power supply

Your next move is to construct a regulated power supply to provide the grid bias voltage you jotted down. A zener diode is highly recommended. Any number of zeners can be connected in series to provide the desired voltage. If you're planning to use two or more tubes in parallel, the grid current will increase in proportion, as will the plate current. After installing the zener-regulated bias supply, recheck to assure that the tube resting plate dissipation is still at the value you've elected to accept.

designing the amplifier

With your odd-ball tube you'll be on your own much more than if you follow some standardized design. Consider tube sockets for example. Many odd-ball tubes require sockets that won't be found in most supply houses. You may have to make your own, which I did for the Eimac 3A200A3 tubes in my amplifier. This job isn't too hard. Sometimes you can find a banana-plug socket that will fit tube prongs. I once used such components to make a socket for a Western Electric 212-E tube. Keep in mind that filament contacts must carry high current; therefore, the contact surface must be amply large and must make a firm connection.

tune up

For initial tune-up tests, it's best to use a dummy antenna. Only the final part of the test requires the radiation of a signal.

You've determined and set the values of plate voltage and grid-bias voltage. Now you must determine the proper excitation (as indicated by the grid-current meter) and the optimum plate-current loading.

Provide moderate grid excitation — just enough to resonate the plate circuit.

Now increase excitation and try loading the plate circuit. Remember to keep grid excitation to a moderate level. If you know the class-C grid current rating for the tube, keep the current below half this value. Adjust the plate circuit *in small* increments. Turn on the power and make quick *adjustments*, then turn off the power. Continue increasing excitation (but don't exceed the limit previously mentioned). Increase plate loading until the plate current dip becomes quite small. A better means of ascertaining the desired loading is with some device for measuring either the rf power output or voltage. When using this method, continue loading the amplifier until the rate of increase of output power (or voltage) approaches zero. Then back off the excitation about 10 percent.

You may wonder about a procedure that requires adjustments to be made quickly. The procedure is valid. You're adjusting for a maximum (peak) power, which will be reached only by random voice peaks. The duty cycle of such peaks is very small; therefore, they will not overload the tube. Your amplifier must handle these peaks without appreciable distortion else splatter will occur on either side of your desired signal.

on-the-air checks

For this test it's best to seek the cooperation of a station some distance away to avoid cross-modulation and intermodulation distortion in the receiver. Select a time and frequency when interference will be minimized. Also, try to have the cooperating station operated by someone who knows the difference between a clean ssb signal and one that is badly distorted. Do *not* depend upon a station having an oscilloscope attached to its receiver. Engineers who work with commercial ssb equipment have been quoted as saying, "Any distortion that can be detected on an oscilloscope already has gone far beyond the limit of tolerance." The instantaneous splatter that results from intermodulation products on the high-amplitude peaks of voice modulation can be detected only by listening on adjacent frequencies or by the use of an extremely expensive real-time spectrum analyzer.*

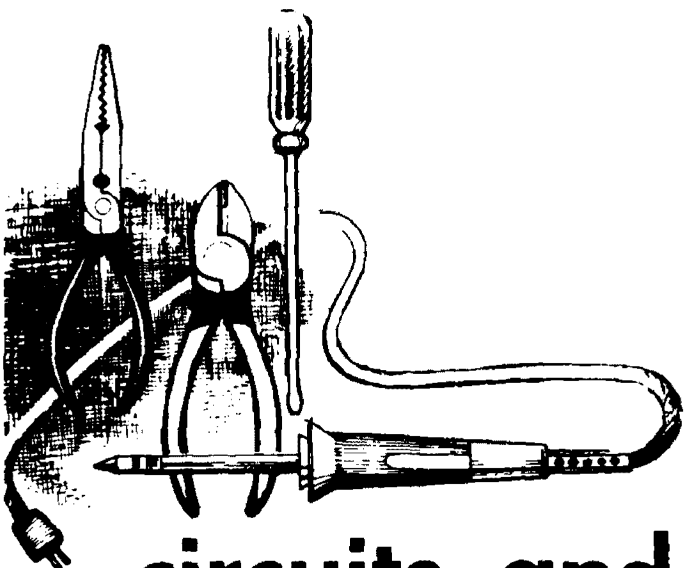
An honest and well-qualified listener, equipped with a conventional a-m receiver with a front end reasonably free of cross-modulation and intermodulation and operated with the agc disabled, af gain high, and the rf gain set to give a moderate signal, can provide meaningful information if he tunes slightly to either side of your signal as you talk normally. Ask the operator to listen for the buckshot effect that results from nonlinear operation of an amplifier. If such distortion shows up only on voice peaks, you're probably overdriving the amplifier. Hold the amplifier excitation low enough to pass the test. Your signal may not kick S meters as high, but you'll have better relations with other amateurs and the FCC.

Linearity is not something that's built into a tube. In addition to the tube characteristics, linearity depends upon external factors such as grid bias, grid excitation, plate voltage, and plate loading. These factors are mutually dependent, so it is not remarkable that true linearity is difficult to approach. Note that you can approach it but never achieve it.

Don't hesitate to juggle grid excitation and plate loading to obtain the best possible signal. Usually the plate loading must be very heavy. As a last resort, try a different value of grid bias remembering that this bias always must come from a source of low internal resistance (strapped-down supply). Also bear in mind that linearity is more nearly approached when the tube bias point is such that the tube's rated plate dissipation is not exceeded.

ham radio

*However, consider the article by Marv Gonsior, W6VFR, in the March, 1972, issue of *ham radio*. Marv shows how ssb signals can be evaluated with a monitor scope connected to a receiver when the receiver bandwidth is appropriately modified to pass essential information. editor.



circuits and techniques

ed noll, W3FQJ

emitter-coupled logic

I introduced the basic types of digital IC circuits in an earlier column.[1] One of these types was the emitter-coupled logic (ECL). These are often referred to as MECL types which is a Motorola terminology.[2] There are several categories of MECL types according to speed of operation. The family identifiers are MECL-I, MECL-II, MECL-III and MECL-10,000.

The identifiers have to do mainly with the two electrical characteristics of toggle rate and gate propagation delay. The toggle rate is the frequency with which the logic activity can be made to changeover in one second. For example, if the family has a toggle rate of 30 MHz, a multivibrator of that class could produce 30 million output pulses per second.

A digital circuit requires a certain amount of time to changeover from one logic to another. Consequently there is a certain delay between the time of the input signal and the change of logic at the output. Such delay is defined as a propagation delay and is, in effect, a measure of the speed with which a digital IC circuit can be made to function. Propagation delay is usually measured in nanoseconds (a nanosecond is 10^{-9} second).

The MECL families are classified as follows:

MECL-I – 30 MHz toggle rate and 8 nsec propagation delay.

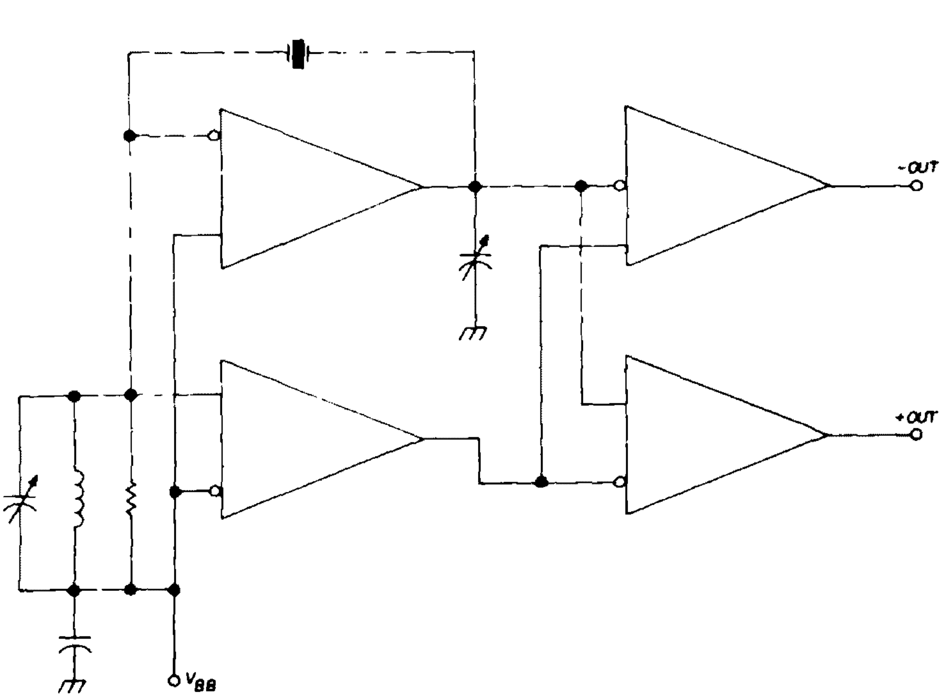


fig. 1. MECL III oscillator using an MC1692.

MECL-II – 70 MHz toggle rate and 4 nsec delay.

MECL-III – toggle rate above 200 MHz and 1 nsec delay.

The MECL 10,000 is a variation of the MECL-III type that fits into large systems and popular circuit wiring methods with greater ease. Propagation delay is 2 nsec associated with an edge speed of 3.5 nsec. Edge speed has to do with the rise and fall times of the pulses. If full benefit is to be derived from high-speed digital circuits, their associated signal input edges must also have a fast time. The maintenance of adequate edges, free of ringing and distortion, imposes strict requirements on interwiring. These requirements are relaxed by the MECL 10,000 designs.

Although the above figures are basic to the various types, it must be pointed out that in each family there are modifications and "improved-upon devices" that are better than the stated figures. For example, MECL-II types are now available with toggle rates of 120 MHz or 180 MHz.

The high-speed ECL types are becom-

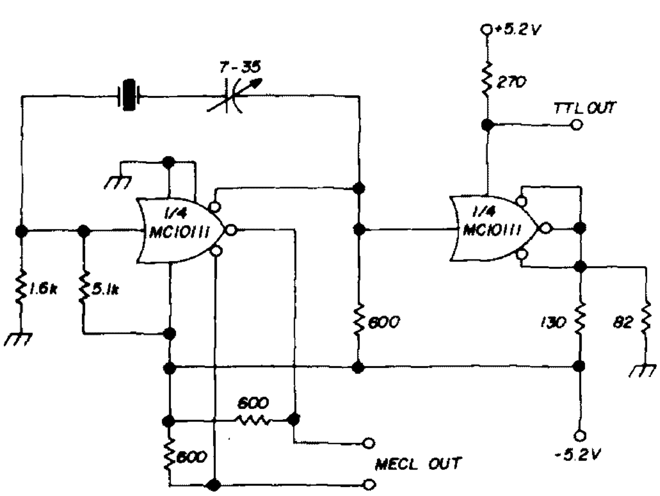


fig. 2. MECL 10,000 oscillator using a MC10111.

ing more prevalent in amateur test gear. Presently they cost more than the popular TTL types but operate at higher frequencies and higher speeds. Often their application as associated with TTL types bear the responsibility for the high speed activities in a given system. In a scaler, for example, an MECL might count down a very high frequency signal to a lower fre-

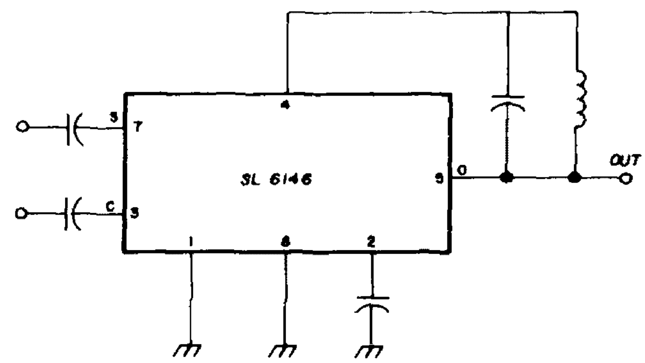


fig. 3. Basic pin-out circuit of a double-balanced modulator using an SL614C.

quency one that can be handled by TTL circuitry. Typical toggle rates for the 7400 TTL series are in the 15- to 30-MHz class. ECL devices are available that can act as a bridge between ECL and TTL types.

high-frequency oscillator circuits

A typical MECL-III crystal oscillator circuit is shown in fig. 1. Its toggle rate is so high that it will operate in oscillator circuits beyond the highest practical crystal frequency. Your high-frequency limitation, in this case, is not the device but the crystal itself. Remember that the output is a square wave, and if not lost in the output coupling system, strong harmonics are present in the uhf and low-frequency microwave spectra.

The circuit of fig. 2 uses an MECL-10,000 type and will operate up into the vhf spectrum. Circuit plan is such that either MECL or TTL outputs are available. The TTL output is such that it can drive a TTL input device or counter. It is said to provide the proper interface between MECL and TTL.

frequency multiplication and division

To some extent the doubly-balanced modulator linear IC has been overlooked by the radio amateur as a means of

frequency multiplication and division. An especially mind-jogging article appeared in the January, 1972, issue of *Electronic Equipment News* (British) stressing the fact that these devices can be used in circuit arrangements that provide more than integral multiplication and division.[3] The device type used was the Plessey SL640 and SL641. The SL640 and SL641 can be made to multiply and divide over a frequency range of 100 Hz to 100 MHz. No doubt similar results can be obtained from the RCA CA3050, Signetics and Motorola MC 1596, and other types made in the United States.

The basic circuit of the device, fig. 3, shows signal and carrier inputs and a tuned output. The tuned output, of course, minimizes the generation of fundamental and other undesired components. In the usual application of a double balanced modulator there are signal and carrier input signals. In the device these two components cancel but mix and produce either a sum or difference frequency (fig. 4A) at the output (F1 + F2) or (F1 - F2). When the double balanced modulator is used as a multiplier or divider, only one input signal is necessary, figs. 4A and 4B. Either the input signal is applied to both device inputs or the second input is derived from the output.

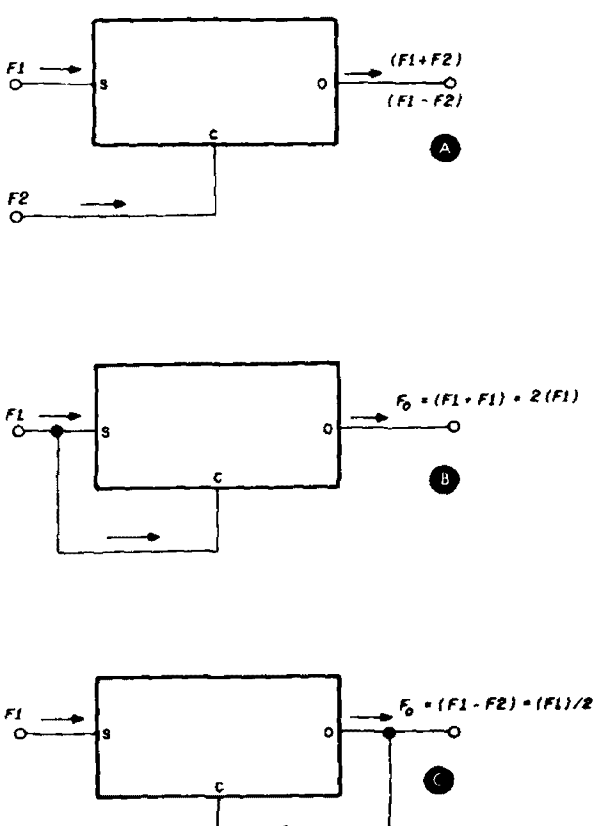


fig. 4. Balanced modulator inputs and outputs. A is the basic circuit, B is the basic doubler and C is the two-to-one divider.

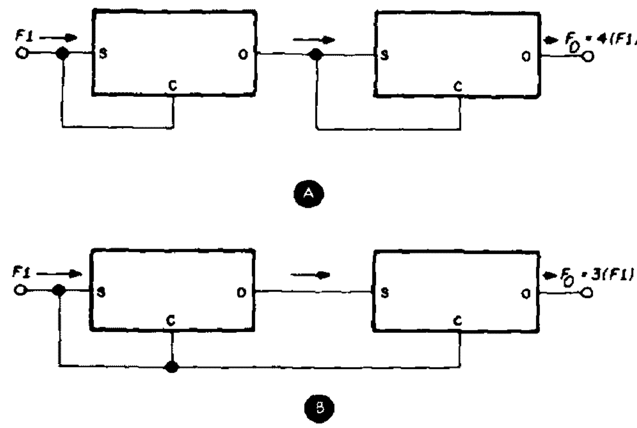


fig. 5. A frequency quadrupler and tripler.

The B circuit operates as a doubler; the C circuit, a 2-to-1 divider. The sum component at the output of circuit A is of course (F + F) or 2F. In B circuit the signal input is Fi and the carrier input is Fo. The output is tuned to the difference frequency, therefore:

$$Fo = Fi - Fo$$

$$2Fo \approx Fi$$

$$Fo = Fi/2$$

Two doublers can be connected in cascade to obtain a multiplication of 4, fig. 5A. Output of first modulator is 2F or (F + F). Output of the second multiplier is 4F or (2F + 2F).

If the input signal is also made the carrier input signal of the second modulator, the balanced modulator pair operates as a tripler, fig. 5B. Output of the first modulator is again 2F while the output of the second modulator is 3F or (2F + F).

A division by four can be obtained by cascading two 2-to-1 dividers, fig. 6A. Odd division using balanced modulators requires feedback paths like the odd-

divider designs of digital integrated circuits. In the case of the balanced modulator, the desired 1/3 output is removed between stages, fig. 6B. Additionally a

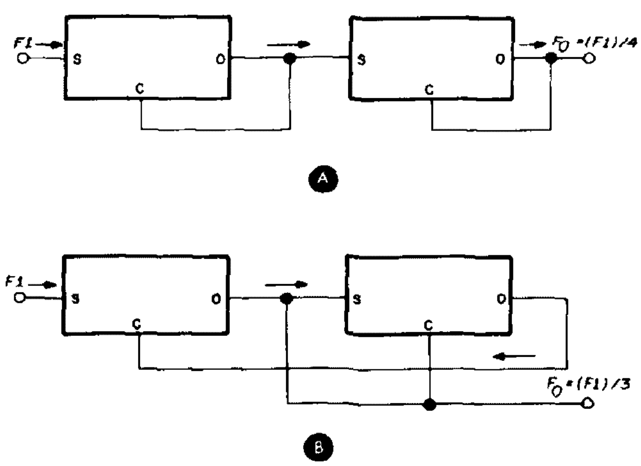


fig. 6. Frequency division by four and by three.

2F/3 component is available at the output of the second stage.

Three, four or five stages can be used to establish various integral and fractional multiplications and divisions. One can anticipate the use of both linear and digital multipliers and dividers in various types of amateur gear including transmitters. Much frequency processing and even modulation can be expected to occur at the low power levels of these circuits. Pulse waveforms often facilitate processing and it is no great problem to convert square waves to sinusoids. This is demonstrated in an earlier experiment where a 7-MHz crystal was used to generate a 3.5-MHz sine wave using a digital two-to-one divider.[4] I am experimenting with these devices and I hope to bring you some practical circuits before the end of the year.

waveform generation

Additional function generators in the form of monolithic chips are being made available. A very appealing one has been developed by Exar Integrated Systems, Inc. In a recent article by Allan E. Grebene, the unusual versatility of such a device was stressed.[5] This 16-pin in-line device can generate sinusoidal, triangular, square, sawtooth, ramp and pulse waveforms. These various waveforms can be amplitude modulated (double-sideband or suppressed carrier), frequency modulated or a combination a-m and fm modulation. The unit can be used to generate a sweep waveform or a tone-burst signal. Regular CW on-off keying is possible as well as FSK or PSK keying. Groups of these waveforms can be made available simultaneously. Here is a possibility for a transmitter design that also includes the generation of its own test and measurement waveforms as well.

The Exar-205 waveform generator consists of three major sections as shown in fig. 7. These are voltage-controlled oscillator, modulator and buffer amplifier. The oscillator makes available linear and ramp waveforms at its output. Frequency range of sine and square-wave generation is 0.1 Hz to 5 MHz. Triangle and ramp waveforms have a range between 0.1 Hz and 500 kHz. The oscillator frequency can be adjusted with a variable d-c voltage applied to pin 13. An audio or other

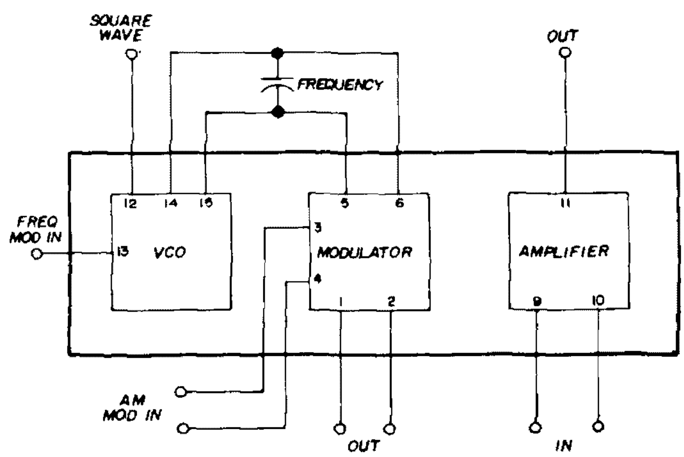


fig. 7. The basic plan for the Exar XR-205 waveform generator.

modulating wave applied to the same pin will produce frequency modulation of the oscillator output.

When a sinusoidal or triangular wave output is desired, the output of the oscillator is applied to the input of the modulator. This wave can be amplitude

modulated by applying an appropriate signal to the second input of the modulator. The modulated or unmodulated sine wave or triangular wave can be taken off at the output of the modulator.

A buffer amplifier is also included in the chip. The signal at pins 1, 2, 12 or 14 can be applied to the buffer, and the amplified output can be removed at pin 11.

A practical circuit arrangement is shown in fig. 8. Ramp voltage is removed from the oscillator between pins 14 and 15. Capacitor C1 sets the oscillator frequency. Oscillator signal is applied to the modulator by way of pins 5 and 6 with a sine or triangle wave output present across pins 1 and 2. The wave-shape of the output is determined by the setting of the potentiometer connected between pins 7 and 8.

Amplitude modulation results when the appropriate modulating wave is applied to pin 3. Potentiometer R2 sets the dc bias voltage at pin 4, regulating the magnitude of the output signal.

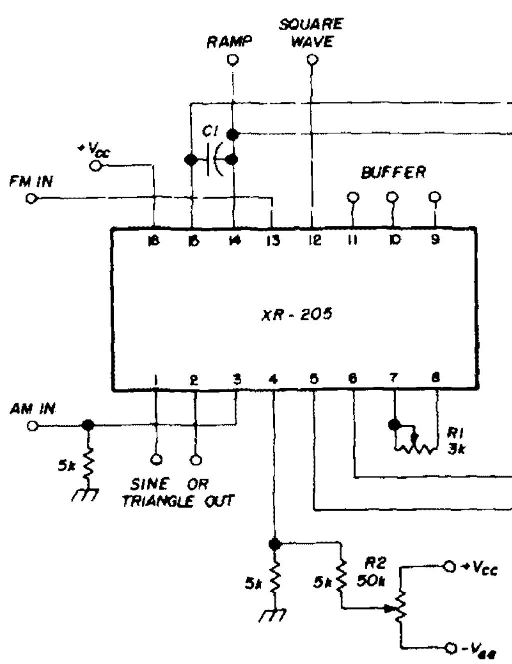


fig. 8. A practical function generator.

six meter QRPP

The Siliconix 2N5912 dual-differential amplifiers perform well in push-pull oscillator circuits because of their uniformity. The device will operate as an oscillator to 100 MHz and above. Total device dissipation is about 500 milliwatts and, therefore, it is no great problem to obtain over a 100 milliwatts output on six meters. The circuit of fig. 9 performs well with the inexpensive EX overtone crystals made available by International Crystal Manufacturing Company. The output capacitor is an 8-pF butterfly connected across a coil made from B&W 3011 stock. The coil is seven turns, center tapped. The output winding is two turns of hook-up wire wound around the center of the primary. It will load the usual 50- to 70-ohm six-meter antenna without the use of an output tuning capacitor. More critical loading to other antenna types can be obtained by adjusting the secondary turns or inserting a series tuning capacitor as shown. The transmitter will operate on other bands with the use of appropriate crystals and balanced output resonant circuits.

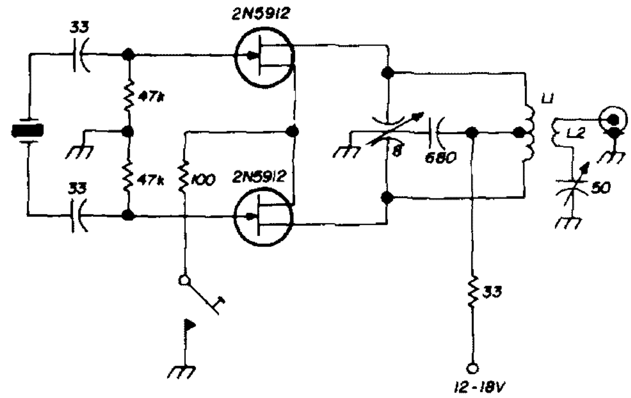


fig. 9. A very low power 50-MHz transmitter.

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1. Edward M. Noll, W3FQJ, "Circuits and Techniques," *ham radio,* April, 1972, page 60.
2. William R. Blood, Jr., "MECL System Design Handbook," Motorola Semiconductor Products, Inc., Mesa, Arizona.
3. James M. Bryant, "Design Aids for Double Balanced Modulators," *Electronic Equipment News* (British), January, 1972, page 51.
4. Edward M. Noll, W3FQJ, "Circuits and Techniques," *ham radio,* July, 1971, page 60.
5. Alan B. Grebene, "Monolithic Waveform Generator," *IEEE Spectrum,* April, 1972, page 34.

ham radio

the ham notebook

electronic keyer paddle

Having built an electronic keyer, I needed a paddle to operate it. Because of cost, poor feel, or my desire not to sacrifice either my straight key or my bug, I decided to build my own paddle. The final model of my paddle cost me nothing, and although it sacrifices a little adjustability, it fits my requirements well and can easily be enclosed in the same box as a keyer, as the paddle only measurers 1 5/8 x 1 x 1-inch.

Fig. 1 illustrates the keyer paddle and gives enough data to duplicate my unit; however, a few comments may clear up some questions. The rubber band allows a one-shot tension adjustment but, by changing the number of loops or by changing the thickness of the band or by varying its position, you can get any tension you want. I was going to use a non-conducting plastic spring but it was expedient at the time to use a rubber band. It worked so well I left it in. You'll probably have to replace the rubber band every so often but its no trouble and it gives you another chance to admire your handiwork. My first one lasted over a year and the second has been going for longer than two years now. It seems to me that when I was a kid we used glycerine on the model airplace rubber bands to extend their life so you might try that. I used one of those small office rubber bands about one inch long and a sixteenth of an inch thick. I send code practice three times weekly in addition to my normal cw operating, so the paddle gets a generous workout.

I used a 6-32 steel machine screw and four nuts for the shaft and bearings. The philosphy here is to get plenty of bearing surface; so large diameter and fine threads are best. However don't go overboard because you might have trouble with foil separation when soldering large nuts to the circuit board. The nuts are ground on one side to allow clearance. A bench grinder, belt sander or such will work here – even a file.

The slots cut in the brass angle stock determine the paddle throw, so be careful not to make them too large. The paddle throw could be made variable in different ways, but generally once throw is set the way you want it, it never is touched again. If you need adjustment, the machine screw can be set a variable distance from the brass angle.

This keying paddle works fine with either a standard or squeeze keyer. My paddle is housed in a 4 x 4 x 2-inch black crackle utility box together with batteries, keyer, sidetone generator and 3-inch speaker. It was built as a present for my wife, but you know how that is.

Kenneth R. Klopf, KL7EVD

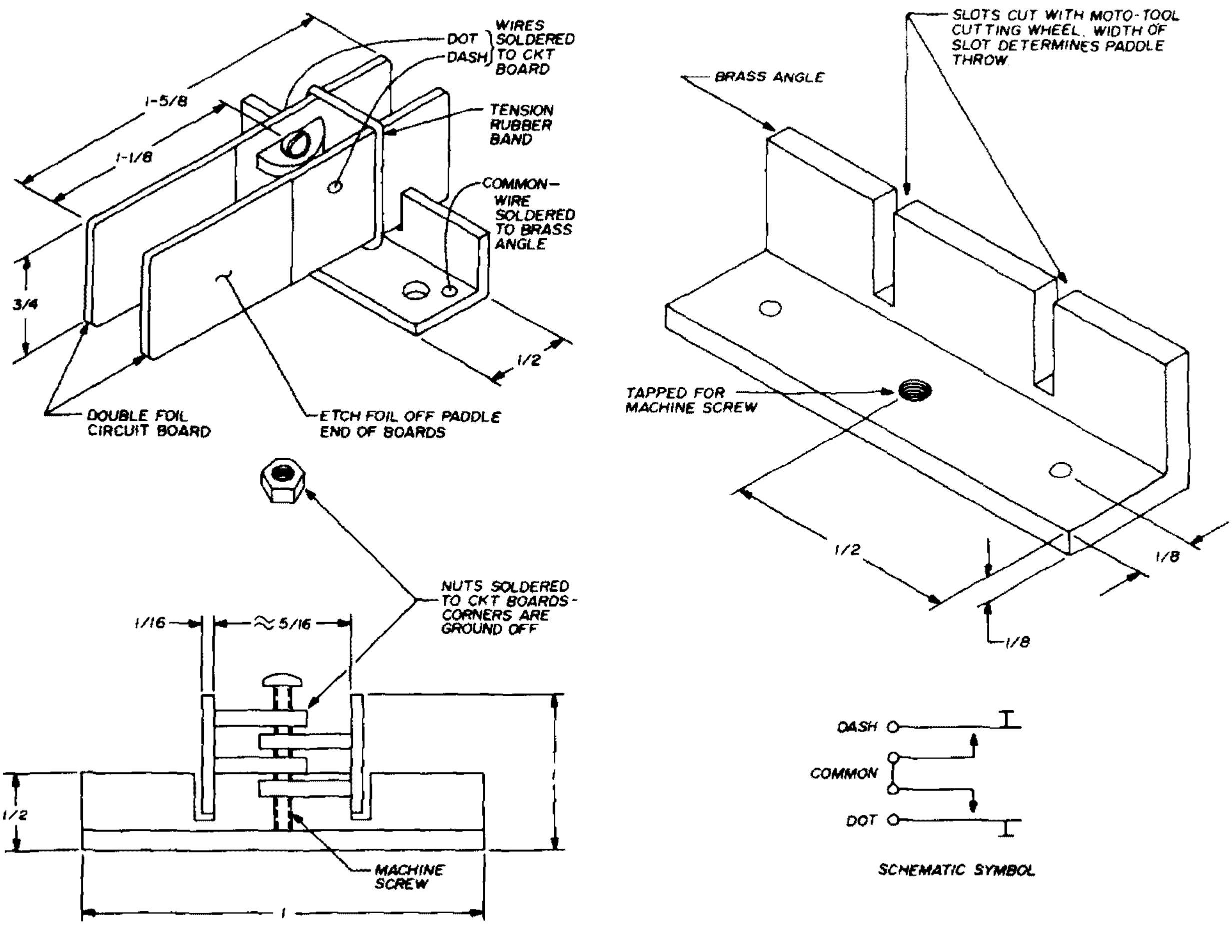


fig. 1. Construction details of the simple keyer paddle.

frequency standard

A stable and highly accurate signal for calibrating counters, meters, or frequency sources can be found in every color television receiver. The local color subcarrier oscillator, at 3.579545 MHz, is crystal controlled and phase locked to the transmitted "color burst" signal from the television station. If the program is of local origin or a delayed tape presentation, this frequency is stable to about one part in 10^8 short term – or one cycle in three seconds. This is more than sufficient for most amateur work and exceeds the capabilities of many inexpensive counters.

If the local station is transmitting a live program from New York such as a national news special, some sports events or some daytime game shows, the color-burst signal will be phase locked to the network rubidium standard and the color TV set's 3.579545 MHz oscillator will be essentially as stable as the standard – typically one part in 10^{12} per day. This resolution is beyond the needs of any but the most advanced or fanatical workers.

Aarne T. Haas, WA7JIK

scanning receiver interference

With the increasing popularity of police-band vhf-fm scanning receivers, a number of amateurs operating on 2-meter fm have had very disconcerting engagements with a new form of interference — scanning receiver interference. SRI appears as an interruption of normal scanning reception of the police band by quite clear and intelligible reception of a local 2-meter amateur.

The cause of the interference is quite simple to understand and, like most interference problems, it is caused by the design shortcomings in the receiver circuitry. These scanner receivers utilize a frontend which uses gated conversion local oscillators capable of scanning some 6 to 8 MHz of the vhf-fm public-service band. Obviously, the rf amplifier preceeding the converter stage in such a radio would have to have a broad response. Unfortunately, some of these are a little too broad and are easily overloaded when operated in a high-strength rf field produced by a 2-meter fm transmitter. This overload in the receiver frontend generates products which look like signals to the rest of the receiver, and the receiver stops scanning and presents beautiful, cleanly-detected 2-meter fm audio — generally to the receiver owner's dismay.

The owner's first reaction is usually to find out "who is interfering with the police and fire broadcasts," and inform him of this fact. This can be an unnerving and embarrassing experience for the unknowing amateur. The correct action to take, though, is the same as for a television-interference problem.

First make sure you are clean. Successful operation of a good-quality police-band receiver while your transmitter is in operation should establish this. The owner of the automatic eavesdropper should then be told *politely* that the problem is one of receiver design. He should be requested to write to the manufacturer to explain his problem and request assistance.

Not all scanning receivers are plagued with SRI. The Digi-Scan, manufactured by Unimetrics, appears to be much less susceptible to this problem than some of the more common scanners such as the Bearcat and the Courier. Of course, good-quality, non-scanning receivers with narrow front-ends are the best choice to prevent SRI. (Adapted, with permission, from *The RaRa Rag*, published by the Rochester Amateur Radio Association.)

Joseph Hood, K2YAH

uhf coax connectors

If you dread installing uhf coax connectors because you have a tough time threading the outer jacket of the coax into the connector, try this. Trim the end of the coax in the usual manner per the *ARRL Handbook*. Daub a drop of silicone grease, about the size of a match head, on the black jacket just back of the exposed braid. As the connector is threaded on, it will pick up the silicone grease and turn on like a nut.

Floyd R. Patten, WØLCP

cold galvanizing compound

After designing and building that antenna to end all antennas, the finishing touch should provide long life. *Cold Galvanizing Compound*, manufactured by Crown Industrial Products Company*, can do the job. This spray-can product is light gray in color and easy to apply. It provides a zinc-rich coating which protects metallic materials from rust and corrosion.

The Cold Galvanizing Compound provides protection in two ways. First, as a long-lived coating of practically pure zinc; and second, through galvanic action. This becomes effective when the surface is scratched or broken — with the presence of moisture galvanic action takes place, but the corrosion of the zinc coating will protect the metal it covers.

Hilary McDonald, W5UNF

*Crown Industrial Products Company, Hebron, Illinois 60034.

new products

ssb and CW transceiver

Hallicrafters introduced a new, low priced, completely self contained ssb and CW transceiver, the FPM-300. It is compactly designed with modular construction techniques for fixed, portable and mobile use for amateur, CD, CAP, MARS, RACES and other utility hf communications services.

The new transceiver provides the user with an extended range vfo (600 kHz) for full-frequency coverage of 80 through 10 meters.

Priced at $595, amateur net, the transceiver features low power drain, conservatively rated 250 W PEP input on ssb and 180 W on CW. The unit is all-American made with American components. It uses glass-epoxy pc boards and over 70 active electronic devices.

Other features of the unit include a large, easy-to-read frequency display which can be interpolated to 1 kHz, combination S/tune meter, built-in vox and semi break-in CW, an IC speech compressor, aalc, 100/50/25-kHz crystal calibrator and a universal power supply for 117/234 Vac, 50/60 Hz and 12 Vdc.

For additional data on the FPM-300, write to the Hallicrafters Company, Amateur Radio-Department PR, 600 Hicks Road, Rolling Meadows, Illinois 60008 or use *check-off* on page 110.

semiconductor lasers

Both Ralph W. Campbell, W4KAE, and Forrest M. Mims have written articles and letters in *ham radio* magazine about laser experimentation. These two men have now come out with a new book on the subject entitled "Semiconductor Diode Lasers." The new book is written in a comprehensive but easy to read style and introduces experimenters and design engineers to the injection laser – one of the most unique and challenging semiconductor devices in electronics today.

The first chapter discusses the history and development of the laser. It explains light-emitting diodes, the injection-laser theory and such lasers as the solid, liquid, gas, plastic, gelatin and space.

Chapters two and three describe the fabrication and the electrical properties of the injection laser. Also explained is coherence, a very significant aspect of laser light.

The remaining chapters deal mainly with circuitry and practical applications. Circuitry encompasses pulse generators, modulators, power supplies, detectors and receivers. Optical systems and viewing devices are also mentioned.

More Details? CHECK–OFF Page 110

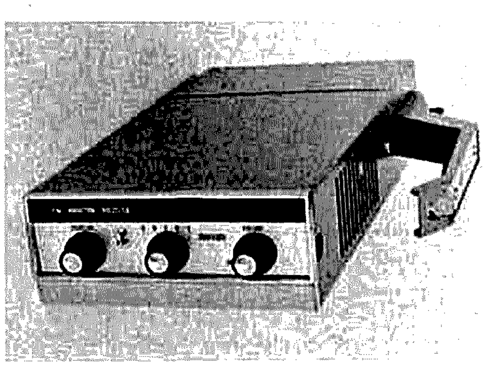
The final chapter covers several of the many applications which are realities and suggests others to come.

This first edition book also contains appendices on laser safety, range equations and addresses of manufacturers. Finally, the book is rounded out by a wide variety of conventional and infrared photographs.

With large-scale, practical light-beam communications still years in the future, this book on laser injection provides valuable research material and is a most interesting book for the experimenter searching for an area in which he can possibly add to man's knowledge.

The book is 192 pages, softbound, and is available for $5.95 from Comtec Books, Greenville, New Hampshire 03048. This new book is published by Howard Sams.

fm monitor receiver



E. F. Johnson has introduced a new non-scanning fm monitor receiver featuring one-at-a-time coverage of five channels. Intended for mobile use in the 150-174 MHz band, the all solid-state unit can be powered directly by a 12-Vdc power source or from an optional ac power supply or field battery pack. The receiver features a dual-conversion circuit with a crystal filter, 5-watts audio output, built-in speaker, true noise-operated squelch, external speaker jack and SO-239 antenna connector. The receiver sells for $119.95.

More information can be obtained from any Johnson dealer, direct from E. F. Johnson, Waseca, Minnesota 56093, or by using *check-off* on page 110.

pulse tone circuit

A miniature thick film hybrid pulse circuit chip is now available from Alpha Electronic Services. The PT-100 pulse chip, when added to an Alpha ST-85 encoder, creates the necessary time delay pulse utilized for pulse or burst tone CTS systems, telemetry, radio controls and selective signalling systems.

The Alpha ST-85J encoder is contained in one thick film chip, the frequency determining network is one thick film chip and the entire unit when combined with the PT-100 chip makes a very small package (7/8" x 1-1/4" x 1/2"). Because of the miniature size, it is especially useful in handheld radio units or mobile units where space is at a premium.

The thick film hybrid technique is especially desirable where exceptional long term reliability is required or where the high failure rate of reeds is a problem.

As a complete pulse tone encoder the ST-85J, PT-100 is available in any frequency from 20 to 3000 Hz. Temperature range is from -40° to +100° C. Current requirement is 4 mA at 12.6 volts, but it will perform within a voltage input range of 6 to 24 Vdc.

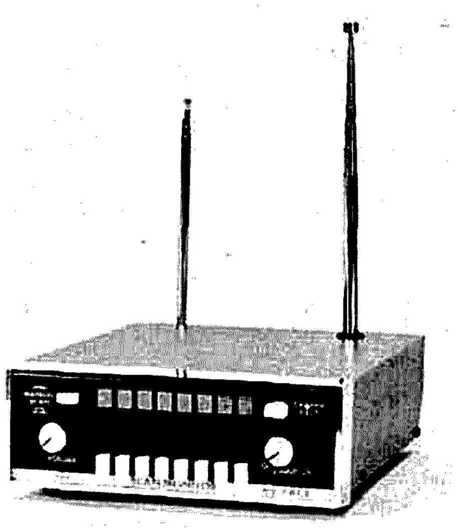
Application engineering assistance is available. Contact Alpha Electronic Services Inc., 8431 Monroe Avenue, Stanton, California 90680, or use *check-off* on page 110.

transistor tester

The model 85 transistor tester saves considerable time in locating faulty transistors. In-circuit tests can be made by placing the contacts of the tester against the printed-circuit board. A tone is heard if the transistor is a functioning unit. The absence of a tone indicates a shorted or open transistor. The unit comes with an input extender cable for adapting to various lead configurations. The battery-powered unit costs $17.95. More information is available from Production Devices, 7857 Raytheon Road, San Diego, California 92111 or by using *check-off* on page 110.

More Details? CHECK—OFF Page 110

three-band scanner



A programmable, three-band scanning monitor receiver covering 25 to 50 MHz, 140 to 174 MHz, and 450 to 470 MHz simultaneously, is now available from the Pace Communications division of Pathcom Inc.

The receiver can be programmed easily for monitoring any combination of eight channels in the high band vhf, low band vhf or uhf frequencies. The unit, designated the SCAN 308, holds up to 16 different channels. With simple switch controls the unit can give visual readout for up to eight channels at one time. It was designed to meet the growing need for multiple channel monitoring by amateur and public safety personnel.

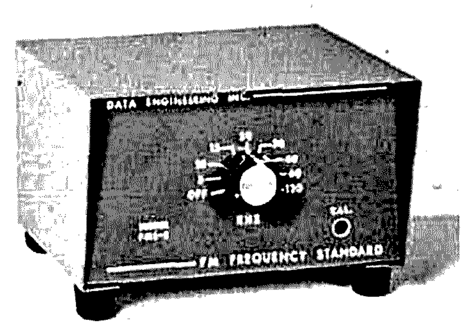
Pace's new SCAN 308 has a wide frontend design so that one model, tuned at the factory, can be easily retuned for extreme field conditions. Technical features of the new model include a unique IC and fet transistor complement to provide versatility of broad band adjustments while still maintaining good selectivity and sensitivity. Rear panel programming switches select the desired combination from 16 internal crystal sockets. No internal wiring need be changed. Front panel control lights, with lock out controls indicate which channel is being monitored.

Built for both 12 Vdc mobile operation and 110 Vac home use, the SCAN 308 is provided with ac and dc power cords, a locking mobile mount, non-slip desk mount, telescoping antenna and built-in speaker. There are provisions for external remote speaker as well as external antenna connections. The wood grain styled metal case makes an attractive desk console.

This unique three band simultaneous scanning monitor sells for $189.95. This model is manufactured in the United States and carries a complete two year service warranty on parts and labor.

For more information consult your local Pace registered sales outlet, write directly to PACE, Box 306, Harbor City, California 90710 or use *check-off* on page 110.

fm frequency standard



Data Engineering is offering a ready-built frequency standard designed specifically for the fm enthusiast. The unit puts out markers at 5, 10, 20, 30, 40, 60 and 120 kHz throughout the 10-, 6-, 2- and 1¼-meter bands.

The unit can be easily calibrated against WWV with an adjustment through the front panel. Completely self-contained, the unit is designed for precise frequency output, cancellation of unwanted markers and usable output above 220 MHz.

Calibrated against WWV and with a built in battery holder for four C cells, the unit sells for $44.50 and carries Data Engineering's five year guarantee. For more information write to Data Engineering Inc., Box 1245, Springfield, Virginia 22151 or use *check-off* on page 110.

More Details? CHECK—OFF Page 110

solid-state electronics

Today, the technician is expected to assume technical responsibilities that formerly were controlled by engineers. As a consequence, the valuable electronics technician, often called an Associate Engineer, must have more than a superficial knowledge of the popular solid-state components now in use. The main objective of "Solid-State Electronics," is to help technicians meet this challenge. George Rutkowski, the author of this new book, not only discusses the fundamentals, but also develops the student's ability to select proper design components for solid-state electronic circuits.

The book begins by explaining common semiconductor materials. Other chapters discuss zener diodes, junction transistors, SCRs, fets and ICs.

A modified programmed style is used throughout the book. Each point discussed is followed by at least one worked example. The student is encouraged to work each sample problem before referring to its solution. The answers to the odd-numbered, end-of-chapter problems are provided at the end of the book. These problems, with the examples, make this book a highly-recommended source for either self-study or classroom use.

"Solid-State Electronics Laboratory Manual" has been written by Jerome E. Olesky to accompany "Solid-State Electronics" to enhance its value as a study course.

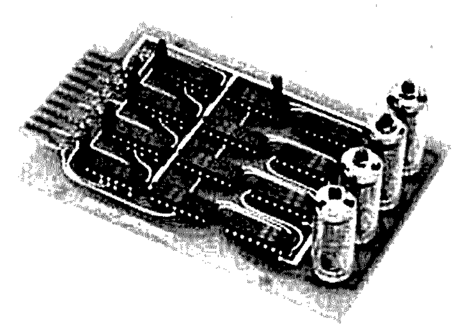
"Solid-State Electronics" is 616 pages, hardbound and costs $15.50. "Solid-State Electronics Laboratory Manual" is 144 pages, softbound and costs $4.50. Both are published by Howard W. Sams and are available from Comtec Books, Greenville, New Hampshire 03048.

theory course

Ameco Publishing Corporation has come out with a profusely illustrated, 448-page course for the first and second class FCC commercial licenses. Broken into 21 lessons, the book is logically arranged so you can study for the second class license and then use the same book to progress towards the first class. The book is quite comprehensive, includes practice tests and deals with over 600 FCC type questions.

The softbound book, catalog number 15-01, is available for $5.95 from Comtec Books, Greenville, New Hampshire 03048.

counter and display modules



A new series of counter and display modules is available from Display Electronics. The CM series modules include a decade counter, latch, decoder-driver and readout for each digit. Standard modules are available with from two to six digits.

All ICs are 7400 series TTL with a typical minimum counting rate of 18 MHz. The modules operate over a temperature range of 0° to 70° C. Lamp test and zero blanking functions are provided. A single 5-Vdc power supply is required for logic and readout. The readout tube is a seven-segment incandescent type with a character nearly one half-inch high. A piece of non-glare polarizing filter material is furnished with each module. Components are assembled on a G10 fiberglass PC board. A rhodium plated edge connector is provided in addition to solder terminals.

The price of a typical four-digit module is $79.00 in single quantity. Custom designed modules are available on special order. Further information is available from Display Electronics, P. O. Box 1044, Littleton, Colorado 80120 or by using *check-off* on page 110.

More Details? CHECK—OFF Page 110

shortwave preselector

The first in a series of specialized products for the forgotten short wave listener has been introduced by Gilfer Associates, Inc. This product is a preselector to enhance weak-signal reception, improve the signal-to-noise ratio in many short wave receivers and virtually eliminate images in single-conversion short wave receivers.

The Model A-20 PreSelector tunes the single range 3.9 to 22.5 MHz. Connected between the receiving antenna and short wave receiver, the A-20 has a noise figure under 2.0 dB and a gain of not less than 18 dB. The gain is variable with a front panel control from near signal cutoff (-40 dB) to the maximum. A slow motion calibrated dial permits setting the A-20 to the peak of its passband (not less than 200-kHz wide at the -3.0 dB points). The A-20 is powered by 117 Vac with transformer isolation for safety. To ensure maximum flexibility, toggle switches are provided for antenna selection and preselector in/out option.

The attractive, compact unit sells for $49.95. More information is available from Gilfer Associates, Inc., Box 239, Park Ridge, New Jersey 07656 or by using *check-off* on page 110.

fm receiver kit

Hamtronics has come out with a new fm receiver kit for the 6 or 2 meter amateur bands or the adjacent commercial frequencies. The kits feature a small size circuit board, low-noise protected mosfet frontend, IC limiter/detector, one-watt audio output stage, narrow-band ceramic ladder filter, posi-

 More Details? CHECK—OFF Page 110

tive acting noise squelch and built-in test features.

The unit is powered by 13.6 Vdc. An optional adapter turns the standard one-channel unit into a six-channel receiver. The builder supplies his own case, controls and speaker and must drill the board himself. This allows custom installation of the board in home-built transceivers and keeps the cost of the kit down.

The receiver board kit is $54.95. The six-channel adapter is $9.95. Discounts are available to clubs and dealers. More information is available upon receipt of a self-addressed stamped envelope to Hamtronics, Inc., 182 Belmont Road, Rochester, New York 14612 or by using *check-off* on page 110.

transistor substitution handbook

The engineering staff at Howard W. Sams has produced the twelfth edition of their popular "Transistor Substitution Handbook." This latest edition is an up-to-date guide providing the reader with over 100,000 transistor substitutions. To guarantee the most accurate substitutions possible, the electrical and physical parameters as described in the manufacturers' published specifications for each bipolar transistor were fed into a computer; then each transistor was compared with all the others. Consequently, transistors which matched within prescribed limits are listed as substitutes.

Section 1 contains substitutions for both American and foreign transistors which are arranged in numerical and alphabetical order. Types recommended by the manufacturers of general-purpose replacement transistors are included at the end of each list of substitutes. Additional data on these general-purpose replacement types — the manufacturer, the polarity, the material (germanium or silicon) and the recommended applications — are reviewed in Section 2.

This handy and comprehensive book is 176 pages, softbound and sells for $2.25. It is available from Comtec Books, Greenville, New Hampshire 03048.

75 cents

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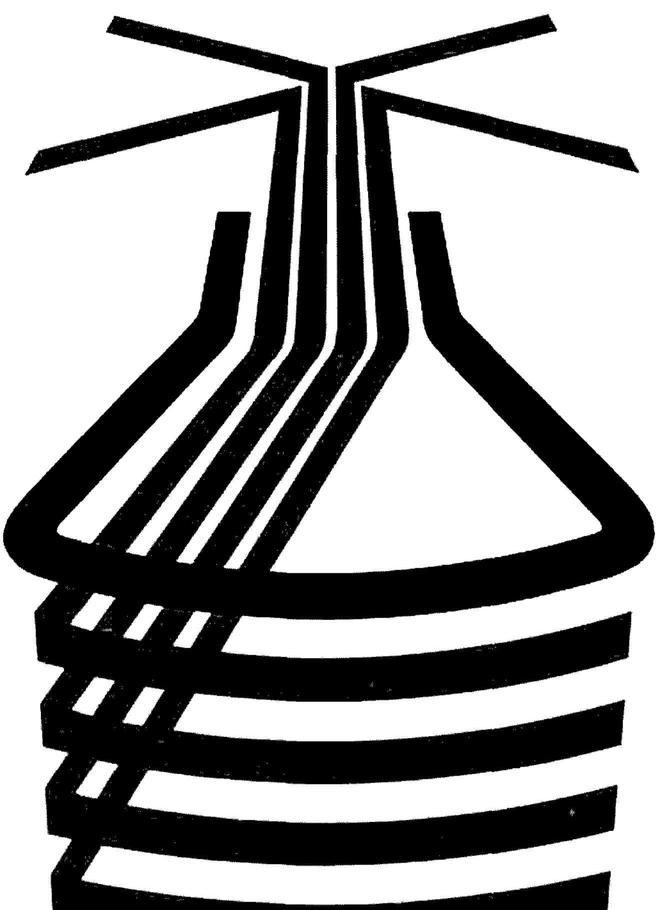
ham radio

magazine

OCTOBER 1972



four-channel spectrum analyzer

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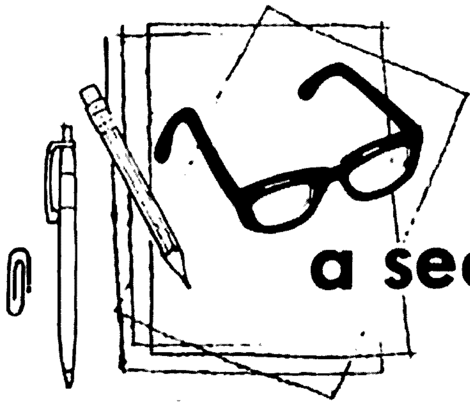


a second look

by jim fisk

The FCC has finally taken action on the so-called fm repeater docket (Docket 18803) which concerns licensing and operating rules for amateur repeaters. If you will remember, in a notice filed in February, 1970, the Commission invited comments on rules proposed for repeater stations. Since that time there has been a lot of discussion between fm operators and repeater groups, with many counter-proposals and comments submitted to the FCC.

On August 29th the Commission adopted several amendments to Part 97 of the rules. Since the present rules do not specifically refer to repeater stations, up until now it has been FCC policy to permit amateur stations to operate as repeaters under the rules applicable to all amateur stations.

Under the new rules, however, a separate station license will be required for every amateur repeater station (beginning July 1, 1973). These stations will be identified by a callsign having the distinctive prefix WR. To qualify for a repeater station license, an applicant must hold at least a Technician Class amateur license, and must submit certain data concerning the technical and operational provisions of his proposed repeater.

An amateur's license, which now specifies the location of his station and his operator privileges, will also include the privileges authorized for his station. At a minimum, the station privilege would be a *primary station.* Various kinds of station privileges may be combined with a primary station license upon submittal of appropriate information.

The remote-control operator may be any qualified amateur designated by the repeater licensee. The new rules permit a licensee to use his own repeater station while he is operating mobile or portable; they also provide for auxiliary link stations to be used when terrain makes multiple-hop control links necessary. The new rules also provide for *wire* remote control.

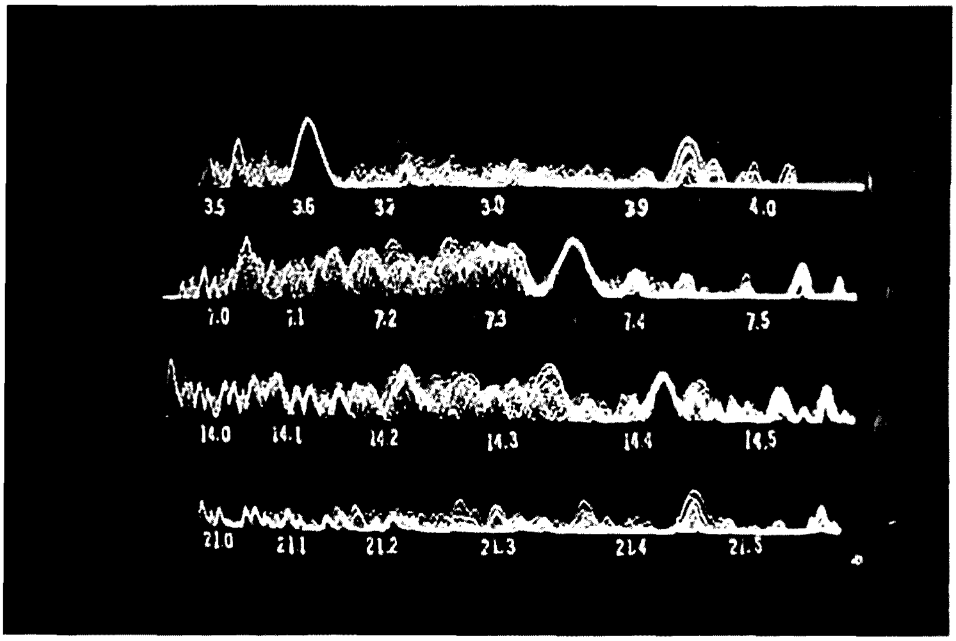
Under the new rules, approximately one-half of each amateur vhf band, and 8 MHz of the 420-MHz band, is authorized for repeater usage, and Technician Class licensees will be permitted to operate in the entire 145- to 148-MHz segment of the two-meter band. The new rules also restrict linked repeater operation, place limits on the effective radiated power from a repeater station antenna, and require the licensee to maintain supervision and control of both the technical and operational performance of his repeater.

The new rules also provide for operation of stations by visiting operators and automatic identification of repeater stations by telephony as well as telegraphy. In addition, they provide for continuous monitoring of remotely-controlled repeaters to prevent interference to communications already in progress on a given frequency.

Although the new rules do not prevent amateur stations from being automatically interconnected to a telephone exchange system, the Commission said that, because of numerous violations of rules regarding interconnection, it may be necessary to examine the use of autopatch facilities and possibly restrict the use of such devices in the amateur service. They also warned that, until new regulations are adopted, interconnection devices must be limited to amateur communication, and may not be used for any type of business communication.

The new rules become effective on October 17th, 1972.

Jim Fisk, W1DTY
editor



four-channel spectrum analyzer

This large-screen, four channel spectrum analyzer will display four different amateur bands at one time

H. F. Priebe, Jr., W9IA, 5040 Wickford Way, Dunwoody, Georgia 30338

The conventional spectrum analyzer displays a portion of the rf spectrum in one continuous sweep. However, for many applications only certain portions of the spectrum are of interest. One such case is the testing and adjustment of frequency-multiplier circuits; another is the monitoring of a particular class of radio service where numerous separate segments or bands are involved. The amateur radio service is just such a service with the popular 80-, 40-, 20-, and 15-meter bands.

Usually, the panoramic adapter[1] or Heathkit *Scanalyzer* has been used in conjunction with the station receiver to provide a number of receiving conveniences. When a-m was popular, the panoramic adapter was exceedingly helpful in net operation. The net-control station could see a signal off frequency when he could not hear him. With the improvements that came with ssb there are far less problems with off-frequency operation. But one of the practical operating problems is that of locating an unused frequency or finding where the action is.

A major shortcoming of the adapter type of panoramic reception is that it is

slaved to the station receiver and the display is centered about the received frequency. You can see what is on the band or part of the band you are listening to, but that is all. And, all too often you know what's happening on the band you're operating on but wonder what the activity might be on another band. That's one of the advantages of the large-screen 4-channel spectrum analyzer. While operating on one band, even during transmission, you can tell what is happening on the other bands.

description

A block diagram of the spectrum analyzer is shown in fig. 1; specifications are listed in table 1. The rf input is connected via an i-f trap to four high-frequency converters, each consisting of a dual-gate, mosfet rf amplifier with agc, a second dual-gate mosfet mixer, and bipolar transistors in an oscillator and switch. An enhancement-mode mosfet in the input circuit switches the antenna so actually only one converter at a time is electrically in the signal path. Converter outputs are 6 to 6.5 MHz, and for simplicity, are broadbanded.

The first i-f, second mixer and second oscillator are electronically tuned with variable-capacitance diodes. Sweep is adjusted to give in excess of 500-kHz dispersion, but even greater range is possible. The two-stage second i-f is at 455-kHz with approximately 10-kHz band-width. A diode detector supplies the rf envelope to the video amplifier. A separate detector is used for agc.

The channel selector and video amplifier consists of suitable synch-pulse shaping circuits, a two-stage binary counter

table 1. Complete specifications of the four-band spectrum analyzer.

frequency coverage	four 500-kHz segments: 3.5-4.0 MHz, 7.0-7.5 MHz, 14.0-14.5 MHz, 21.0-21.5 MHz
sensitivity	1 μV (50-ohm source) provides ¼" deflection; 0.1 V (50-ohm source) provides 2" deflection
resolution bandwidth	10 kHz
display flatness	better than 3 dB on any band segment
sweep frequency	60 Hz sequenced through four segments for frame rate of 15 Hz
spurious signals	none for the four band segments

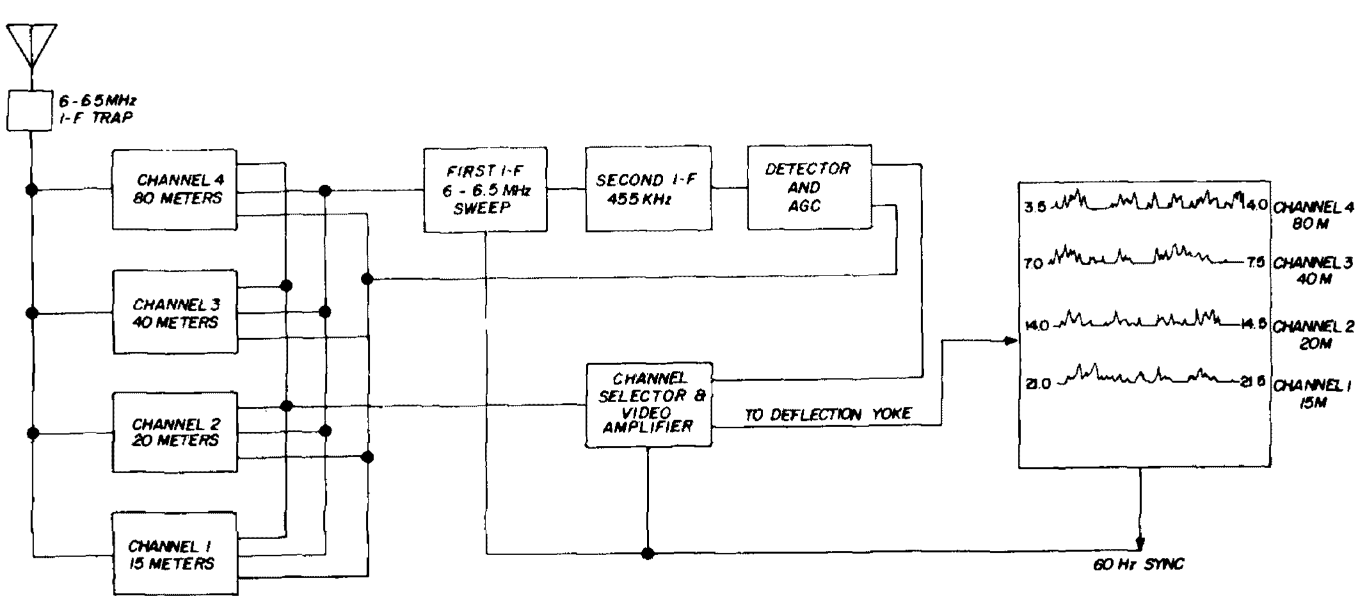


fig. 1. Block diagram of the four-channel spectrum analyzer.

and translator. This system provides the time-division multiplexing of the four converters to the common swept i-f circuit and provides the crt with a four-trace display.

few hundred ohms so the match to a 50-ohm source is not too good, but this has been no problem. The converters are set to the same sensitivity with the rf gain adjustment. However, in actual on-the-air

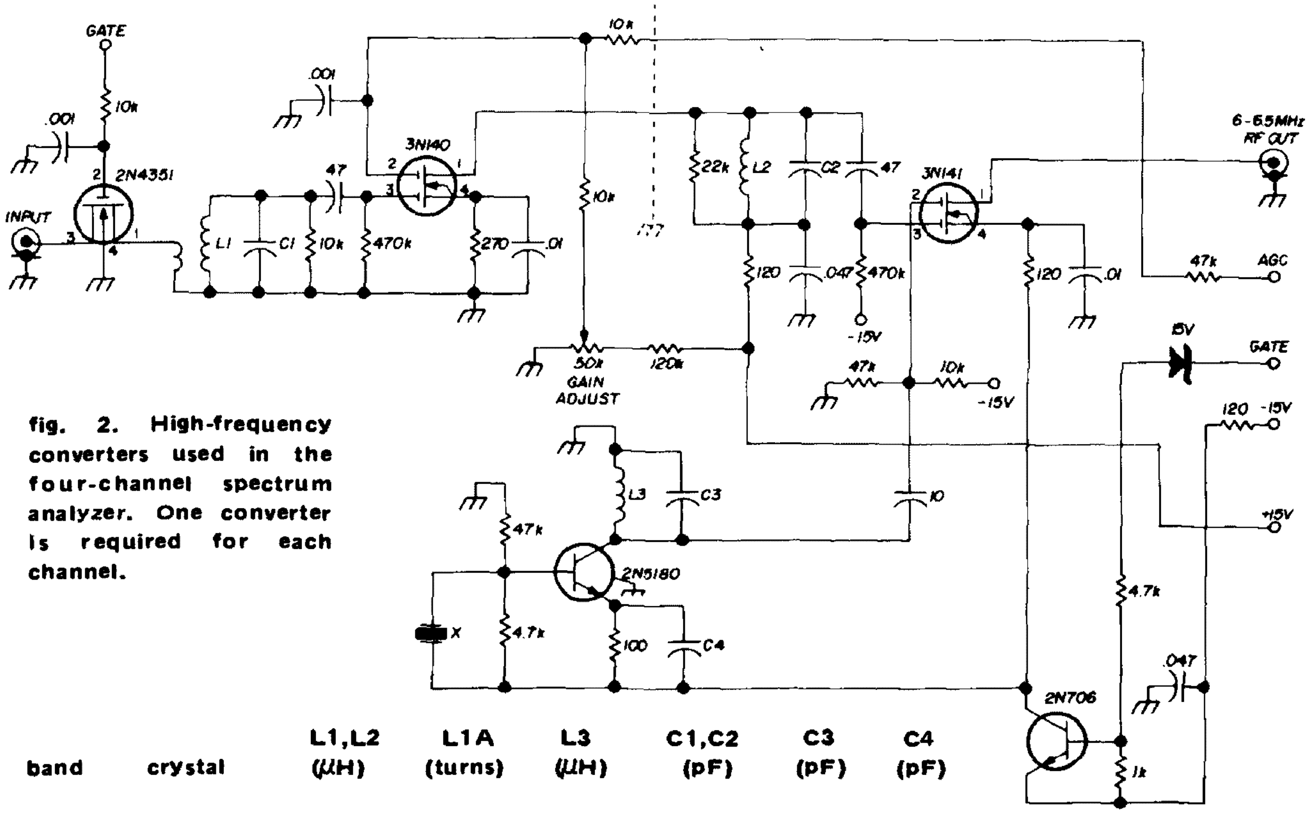


fig. 2. High-frequency converters used in the four-channel spectrum analyzer. One converter is required for each channel.

band	crystal	L1,L2 (µH)	L1A (turns)	L3 (µH)	C1,C2 (pF)	C3 (pF)	C4 (pF)
80	10,005 kHz	16	6	3.2	100	68	220
40	13,505 kHz	8	5	2.5	47	47	200
20	20,505 kHz	4	4	1.5	27	33	82
15	27,505 kHz	2	4	1.1	68	27	68

The display unit is a converted tv set which provides a large picture that is easy to view from across the room. The horizontal sweep is synchronized with the 60-Hz power lines; a pulse from the sweep circuits in the tv set is used to synchronize the receiver.

converters

The circuit diagram of one high-frequency converter is shown in fig. 2. All converter circuits are the same; only the tuned circuits and the crystal are different.

The antenna is connected through a normally-off enchancement-mode mosfet which is turned on. The on resistance is a

operation it is desirable to reduce the gain of the lower-frequency units since 1-microvolt sensitivity at these frequencies produces a very noisy base line. A 5- to 10-microvolt senstivity is more realistic.

The output of the mixer stage is connected in parallel with the other converters and each mixer-oscillator combination is gated through a transistor switch. A 15-volt zener diode is used to shift the level of the channel-select gate pulse to operate the switch with negative voltage supply.

sweep i-f amplifier

The sweep circuits are shown in fig. 3. Rf input from the converters is 6 to 6.5

MHz. The sweep oscillator is on the low side with a frequency of 5.545 to 6.045 MHz to produce a second i-f of 455 kHz. Each of the three tuned circuits has its own variable-capacitance sweep-voltage .01-μF capacitor via a 2 megohm resistor. Each capacitor is discharged by a clamping transistor with an adjustable emitter potential. The emitter potential is adjusted at the high-frequency end while

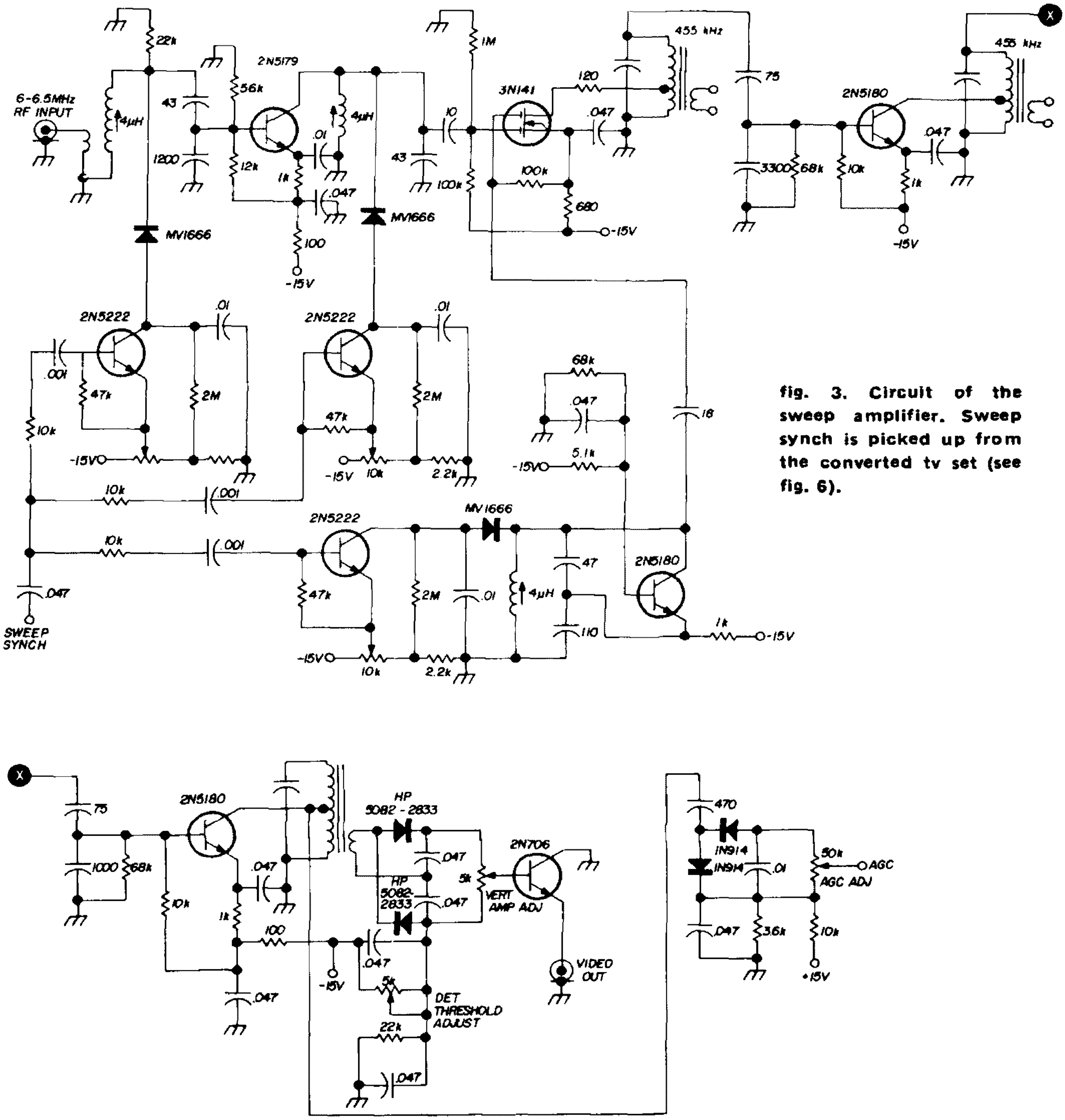


fig. 3. Circuit of the sweep amplifier. Sweep synch is picked up from the converted tv set (see fig. 6).

time-constant components to permit independent adjustment.

The voltage applied to the variable-capacitance diode is developed across a the inductor is adjusted at the low-frequency end. Since the hf oscillator is on the low side of the first i-f frequency, the left-hand side of the crt display is the low

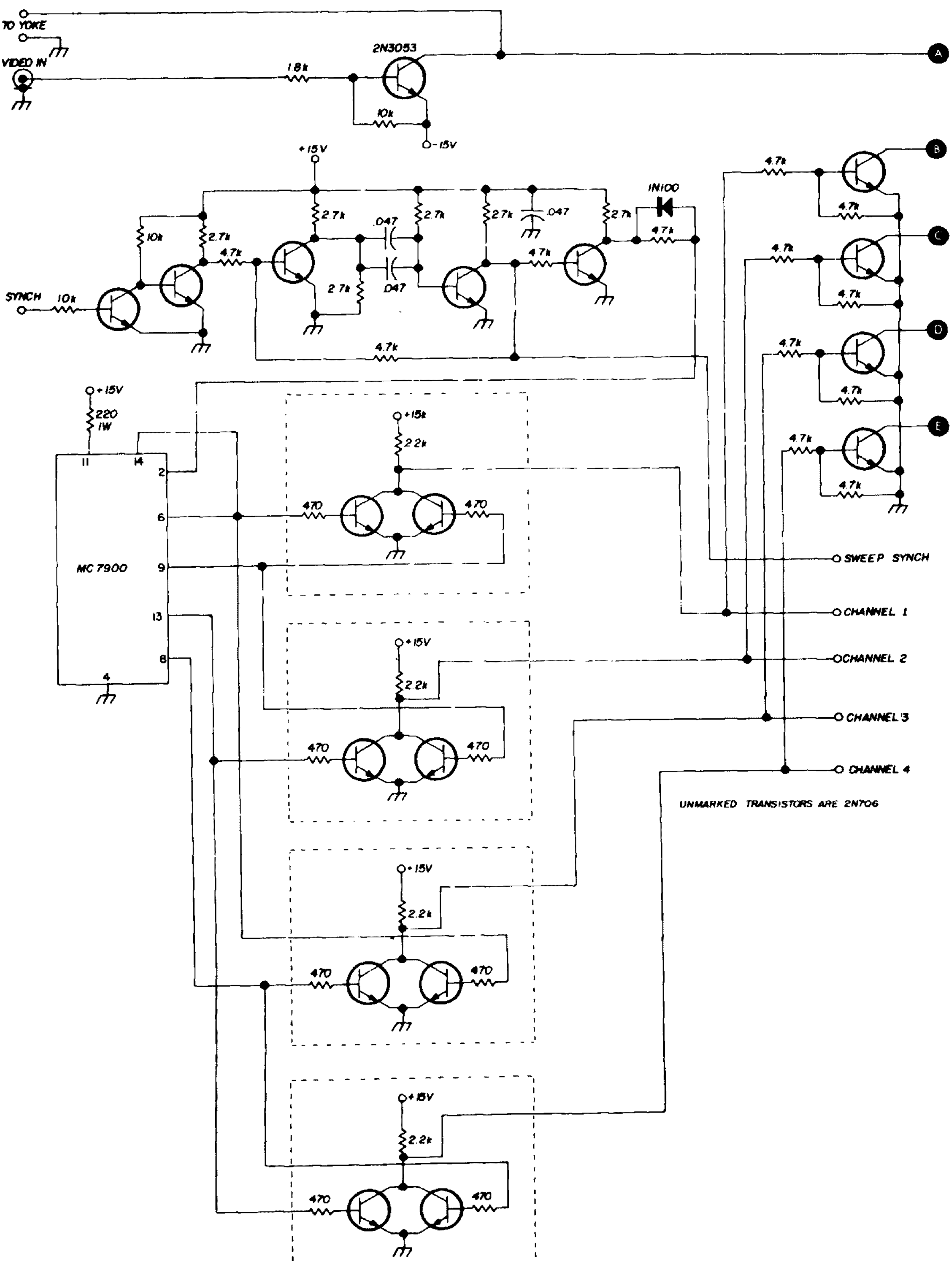
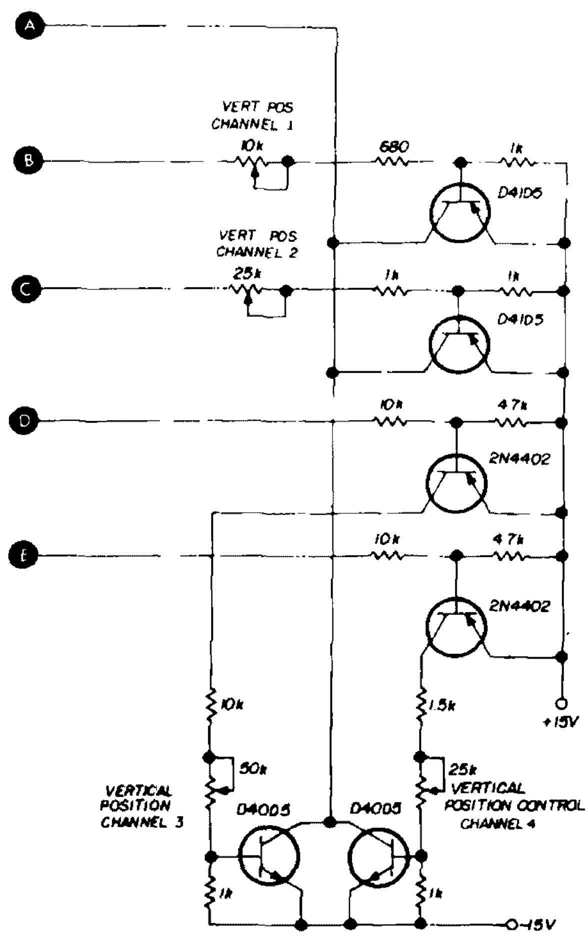


fig. 4. Vertical deflection and channel selector circuits for the spectrum analyzer.



end of the rf input but this corresponds to the high-frequency end of the swept i-f.

The vertical gain adjustment permits use of various size cathode-ray tubes since the larger screens require greater deflection current. The detector threshold adjustment is used to limit base line noise and to offset the input voltage threshold of the video amplifier (vertical deflection amplifier).

channel selector

The four channels are obtained with the circuit shown in fig. 4. The 60-Hz synchronizing signal from the tv chassis is shaped by an overdriven amplifier and monopulser. A two-stage binary counter, a MC790P IC, and the four 2-input AND gates provide the four-channel gate signals. Each gate signal drives a constant-current vertical position circuit. Constant-current positioning circuits are used to give maximum band width since the high current, low voltage and rather large yoke inductance would otherwise give a long time constant and consequently, narrowband response.

Each trace is positioned by setting the base current to the constant current transistor. Adequate range is provided to handle a crt size of 16 to 21 inches with reasonable variation in transistor current gain.

Two voltages, +15 and −15 volts, are obtained from a single transformer with two rectifiers and regulators as shown in fig. 5. These circuits are conventional.

converting the tv

The portions of the tv receiver which remain are the cathode-ray tube, its high-voltage supply and portions of the verti-

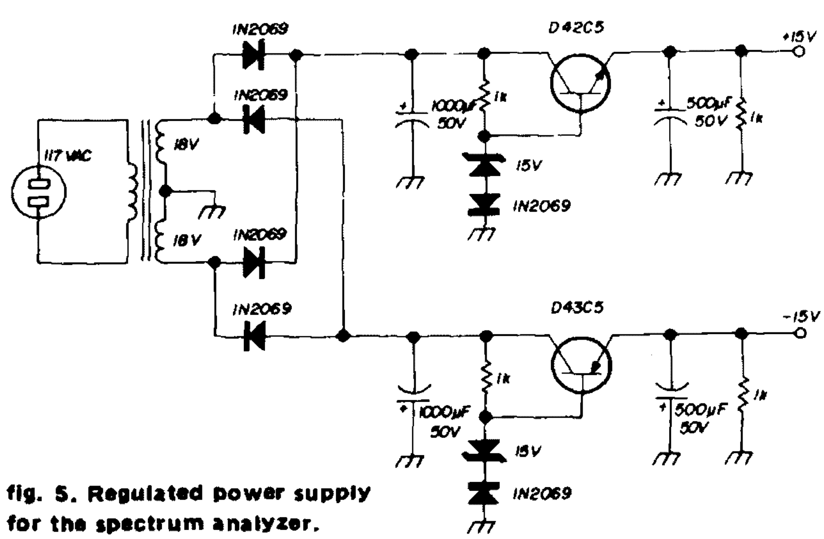


fig. 5. Regulated power supply for the spectrum analyzer.

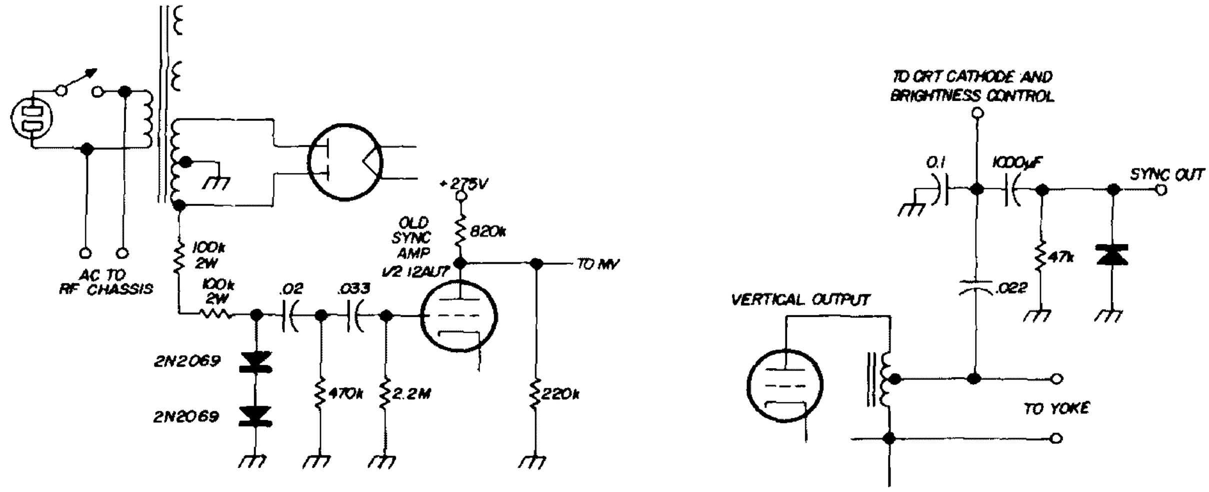


fig. 6. Modifications to the tv set used for display in the four-channel spectrum analyzer. An improved 60-Hz sweep circuit is shown in fig. 7.

cal deflection circuits, as well as the power supply. The conversion involves three steps:

1. The deflection yoke is rotated 90°, and the two leads from the horizontal yoke are brought out to a connector. These are used for the vertical deflection. To insure that the high-voltage flyback supply in the set will function normally, an old yoke, horizontal winding, is connected in place of the yoke wires that are removed.

2. An ac outlet is added on the tv chassis or cabinet and connected to the ac supply in the tv set that is energized through the switch. When the tv set is turned on, power will be applied to this outlet which furnishes power to the panoramic receiver rf unit.

3. A synch pulse is brought out through a shielded lead from the vertical output circuit. These circuit modifications are shown in fig. 6.

With these simple modifications sweep linearity and stability are not the best. Although they work, eliminating the vertical multivibrator and adding the changes shown in fig. 7 are well worthwhile.

When converting the tv set it is advisable to remove all the unused parts, the tuner, the i-f, the audio, speaker, etc. The only functional controls on the tv set are the on-off switch, brightness, vertical linearity and vertical height. However, since

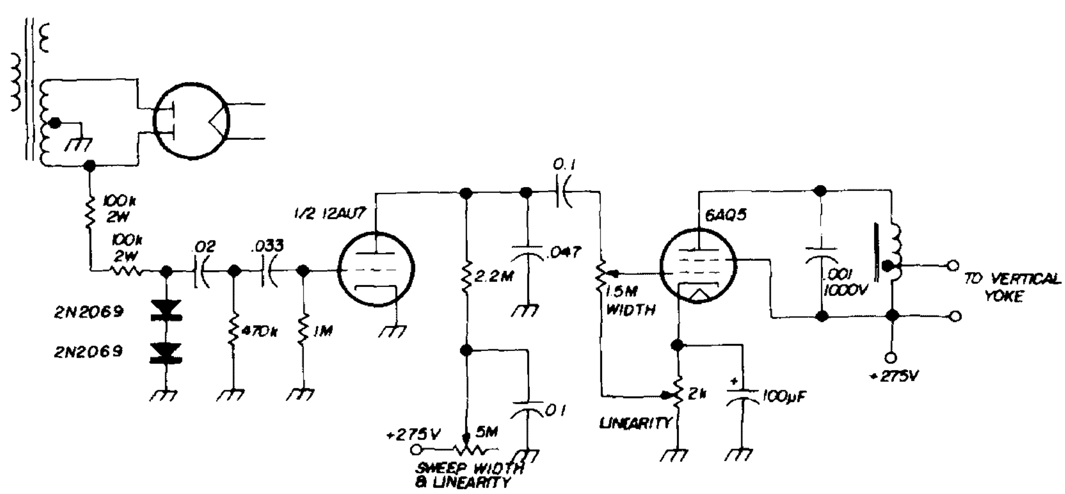


fig. 7. Improved 60-Hz sweep circuit for the modified tv set.

the yoke was rotated 90°, the old vertical linearity and height controls are now horizontal linearity and width adjustments.

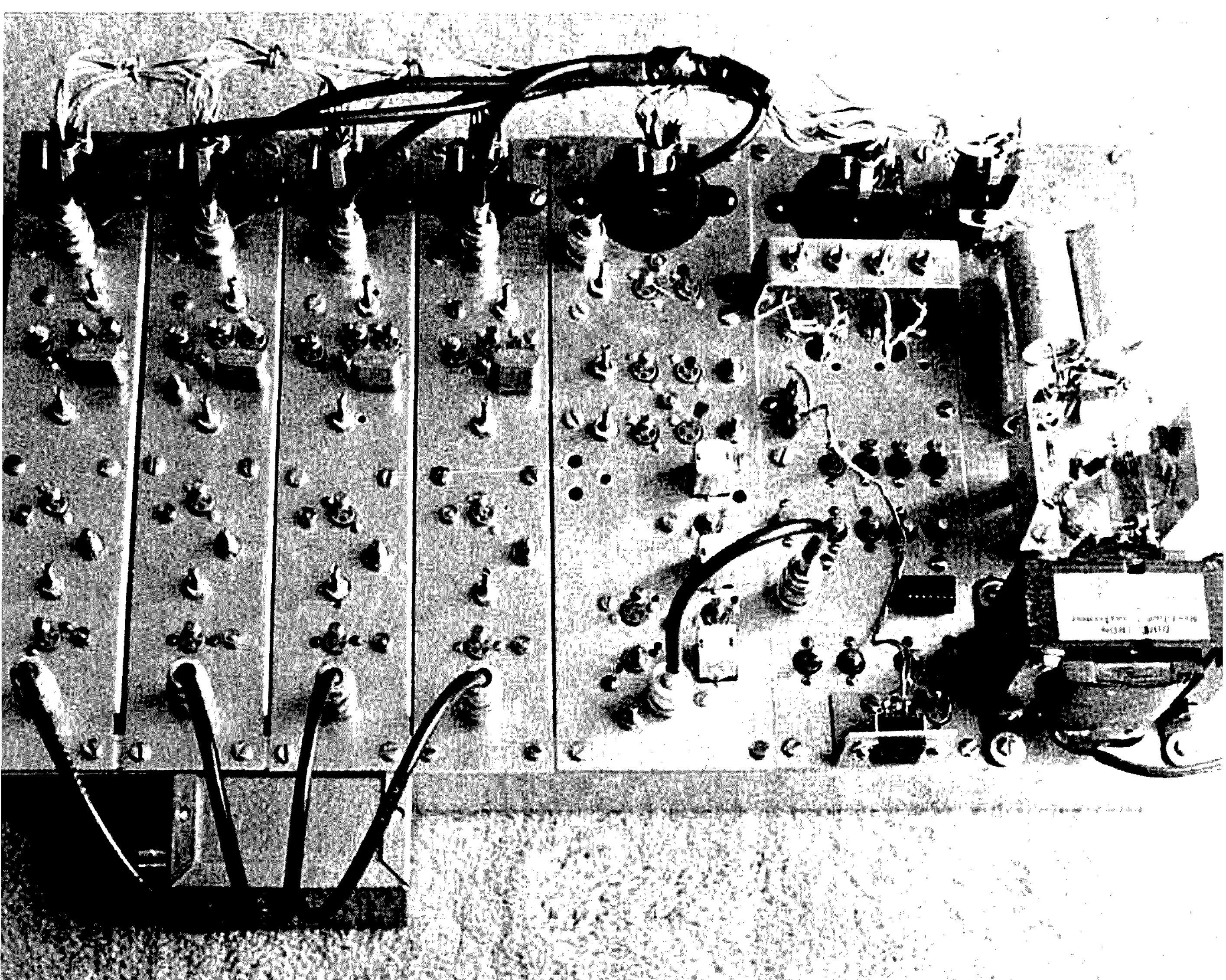
construction

The panoramic receiver, not including the display section, is assembled on seven individual L-shaped subchassis. These are assembled together in a 10- by 17- by 2-inch aluminum chassis. The top of the chassis was cut out and the individual chassis mounted in place of the cutout. The bottom plate is used as this allows troubleshooting the entire assembled unit by merely removing the plate. This could not have been done if the individual subchassis were assembled underneath the chassis, although it would have been a lot easier.

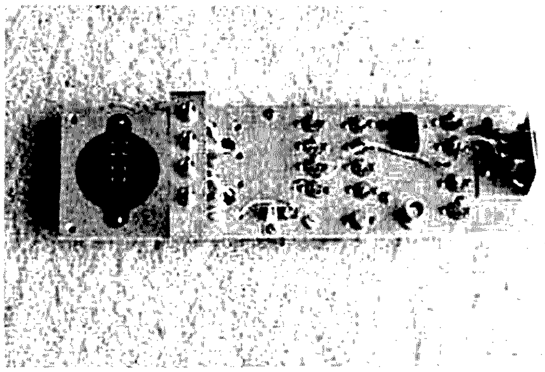
All the subchassis are 2 inches deep by 10 inches long; the rf subchassis are each 2 inches wide, while the sweep, video multiplexer and power supply are each 3 inches wide. All unit interconnections are made through connectors. Rf signals use coax and coaxial connectors and power supply leads and others use Jones plugs.

operation

If you use a large-screen tv set as I did, you will find plenty of room inside the tv cabinet for the receiver circuits. I put the rf chassis in the bottom of the tv cabinet



Complete rf chassis for the spectrum analyzer. Four high-frequency converters are on the left; sweep i-f, vertical deflection, channel selector and power supply are to right.



Vertical deflection and channel selector circuit chassis.

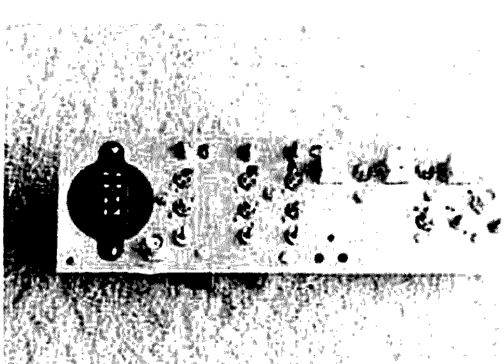
behind the speaker grill. One of the nice things about the console tv is that it doesn't take up any room on the operating table and the large screen can be viewed from anywhere in the radio shack. The unit is excellent for keeping up to the minute on band activities while you are occupied with something else. For the past year I've had mine in operation in the workshop.

The sensitivity of the receiver permits excellent operation with just a short piece of wire for an antenna, although it is designed for a 50-ohm input.

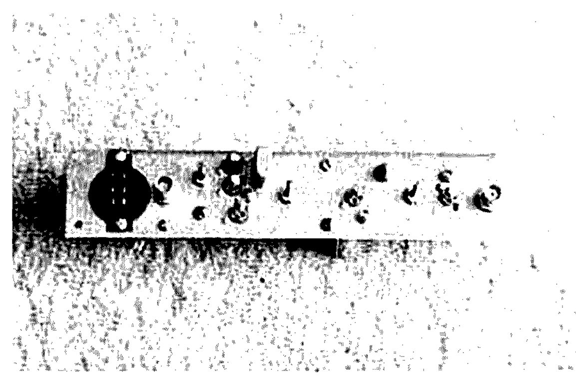
additional uses

The dynamic range is approximately 100 dB. With a 100,000-microvolt signal this results in a 2-inch vertical deflection, while a 1-microvolt signal produces 1/4-inch vertical deflection. The vertical deflection does not give much discrimination between strong signals because of the wide dynamic range.

A very useful application of this multi-band spectrum analyzer is in trouble-shooting harmonic-producing circuits such as frequency multipliers, or when measuring harmonics in amplifiers and oscillators. If a conventional frequency analyzer is used to measure the harmonic amplitudes of a 7-MHz fundamental, the second harmonic is 14 MHz and the third harmonic is 21 MHz, etc. Therefore, each pair of pips is separated by 7 MHz. Since, in this example, we are interested in only three frequencies with a slight dispersion to either side of each frequency, the four band segments are more efficient. With four channels, the fundamental appears on one segment, second harmonic on another, and the third on still another. The total dispersion of the four bands is 2 MHz, whereas with a conventional spectrum analyzer, it would be greater than 14 MHz.



Sweep i-f chassis.



High-frequency converter.

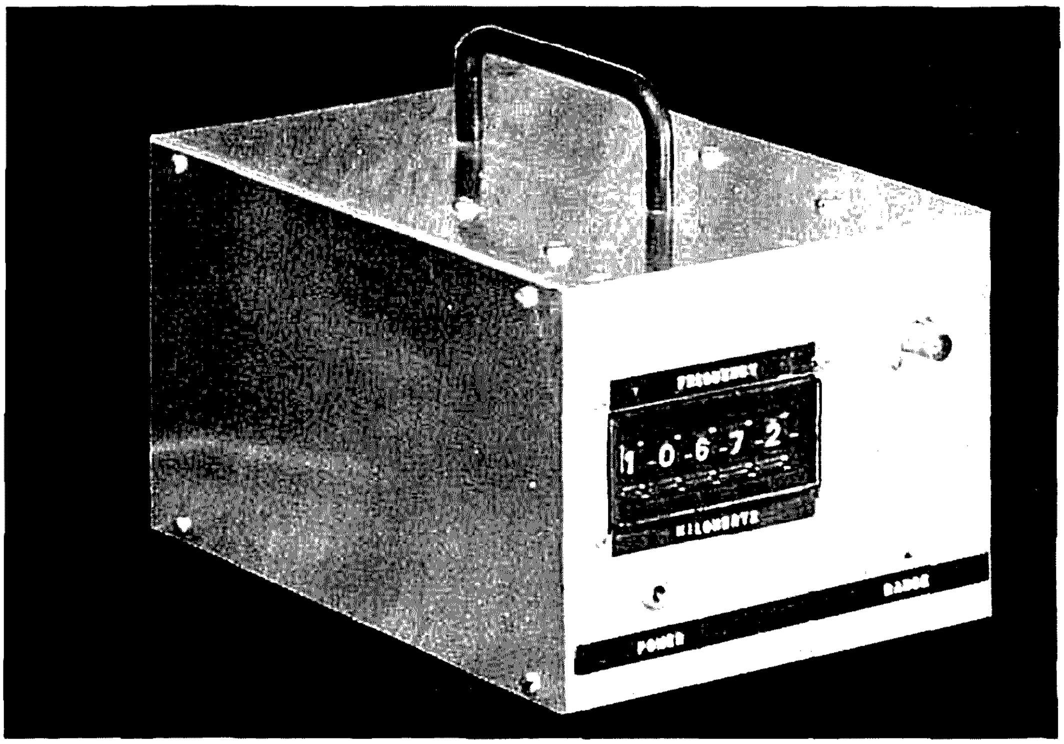
conclusion

If you think this would be a worthwhile addition to the radio shack, you will have to build one because none are being manufactured at the present time. The closest you can come with commercial gear is a general-purpose spectrum analyzer at considerable expense, and you still will not have the desirable feature of selecting the segments of the spectrum you are interested in.

reference

1. H. F. Priebe, Jr., "Build Your Own Panoramic Adapter," *QST*, September, 1954.

ham radio



high-frequency frequency synthesizer

This inexpensive frequency-synthesized 300-kHz to 10-MHz signal generator keeps cost low and performance high through the use of low-cost ICs

Albert D. Helfrick, K2BLA, 115B Linn Drive, Verona, New Jersey

Many amateurs own some sort of radio-frequency signal generator. It is used mainly as a source of accurately known frequencies for receiver or transmitter alignment. Most signal generators found in the ham-shack are capable of ±1% accuracy, at best, which is insufficient for modern radio alignment. For example, 455 kHz on the dial of a signal generator of 1% accuracy would actually be somewhere between 459.5 kHz and 450.5 kHz. This sort of accuracy makes the signal generator of little value for aligning a modern ssb receiver with a passband of only 2.1 kHz.

One of the techniques used to remedy this situation is continuous monitoring of the output frequency with a frequency counter, and periodically adjusting to correct for drift of the signal generator. This technique works quite well, but it is annoying to reset the signal generator constantly.

It would be nice if an electronic circuit could be made to read the counter and automatically make the frequency change. This is exactly what is done by the phase-locked loop synthesizer. The

circuit not only checks the frequency and makes the adjustment, but does it a thousand times per second.

The phase-locked loop frequency synthesizer consists of four basic parts; the voltage-controlled oscillator, the programmable divider, the frequency discriminator and the reference frequency source (see fig. 1).

voltage controlled oscillator

The vco is just as its name implies – an oscillator where the frequency is controlled by voltage. This can be achieved in several ways. A varactor diode may be placed across a vfo tank circuit so that the varying diode capacitance with bias voltage changes shifts the vfo frequency. This type of oscillator usually has a limited range and requires considerable bandswitching or manual tuning.

However, integrated circuits have been designed especially for frequency synthesizers; they overcome the narrow tuning range and provide a more than three-to-one frequency range. The Motorola

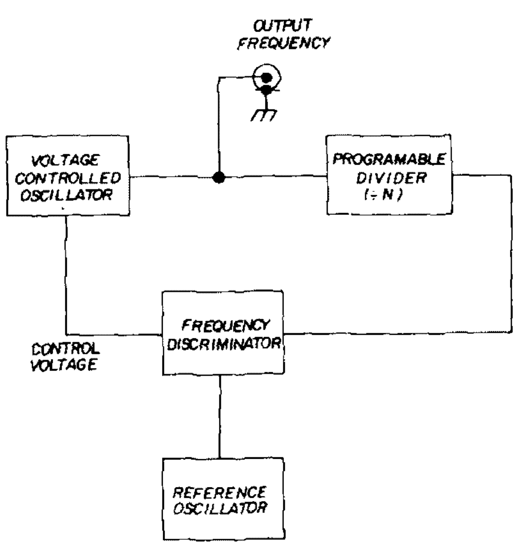


fig. 1. Block diagram of the phase-locked loop frequency synthesizer. Unit goes from 300 kHz to 10 MHz in 1-kHz steps.

MC4024 voltage-controlled multivibrator operates from a single 5-volt supply to frequencies over 30 MHz. A graph of frequency vs voltage for the Motorola 4024 is shown in fig. 2.

programmable divider

The output frequency is determined by the programmable divider. The divider produces an output frequency which is

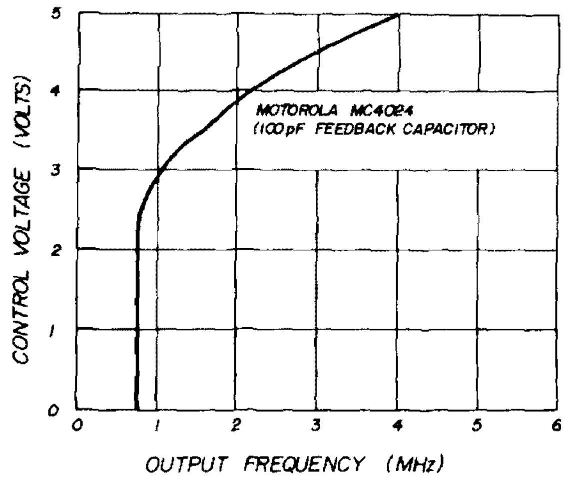


fig. 2. Output frequency vs control voltage of the Motorola MC4024 voltage-controlled multivibrator IC.

the input divided by some N, which is selected by the frequency switches. For example, if the input frequency is 2 MHz, and the switches are set to N = 2000, the output frequency would be

$$f_{out} = \frac{f_{in}}{N} = \frac{2\text{ MHz}}{2000} = 1000\text{ Hz}$$

The popular SN74192 decimal counter IC was chosen for this circuit because of its low cost, simple operation and counting speed. A one-shot multivibrator was required to insure proper resetting of the counters. Sometimes the programmable divider is referred to as a frequency-to-frequency converter.

reference frequency

The reference frequency is generated by a 1-MHz crystal oscillator which is divided by 1000 to produce a 1-kHz square wave. Three SN7490s are used to divide the 1-MHz signal to 1000 Hz. A SN7400 gate is used for the crystal oscillator/buffer.

frequency discriminator

The frequency discriminator is a com-

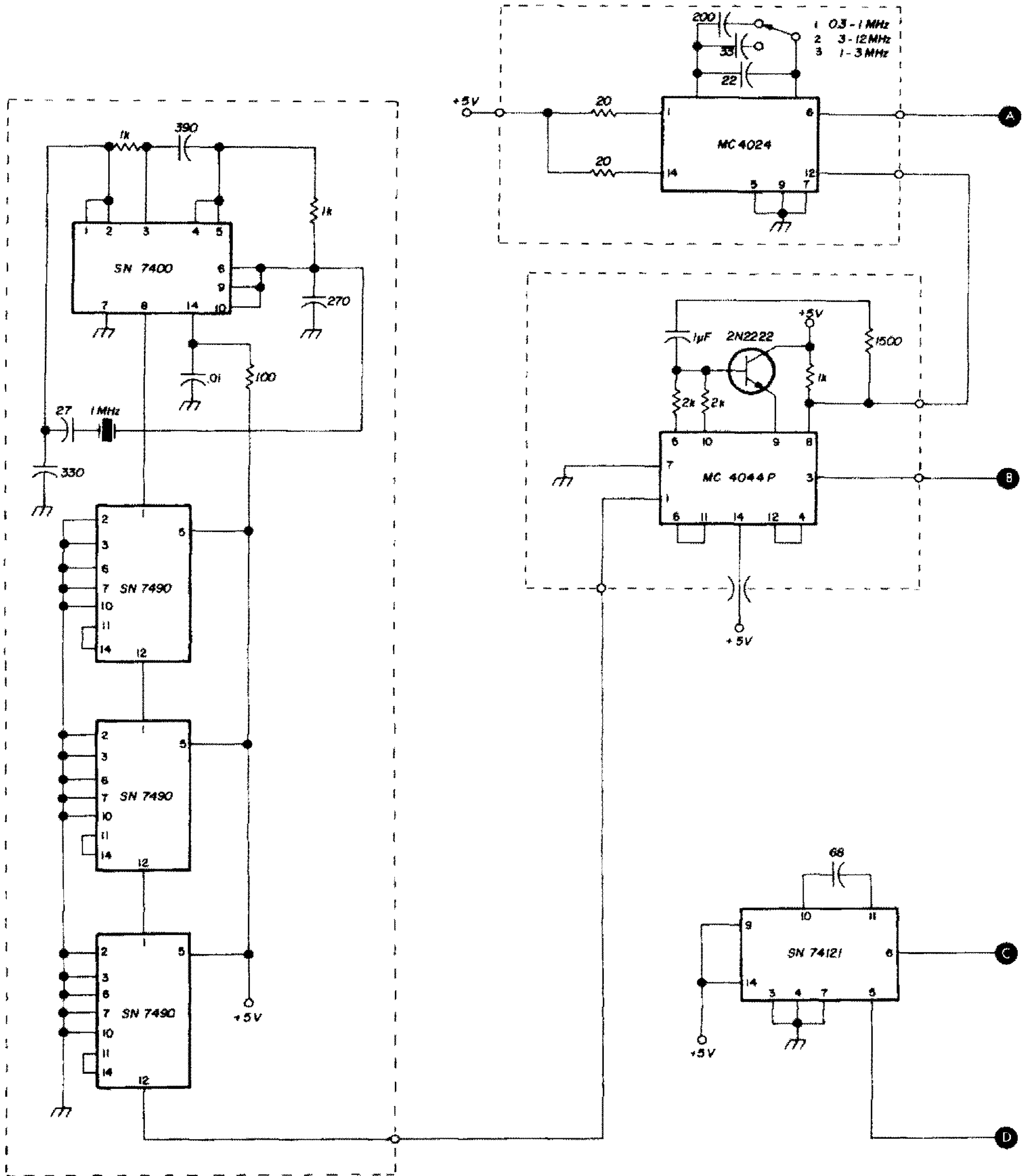
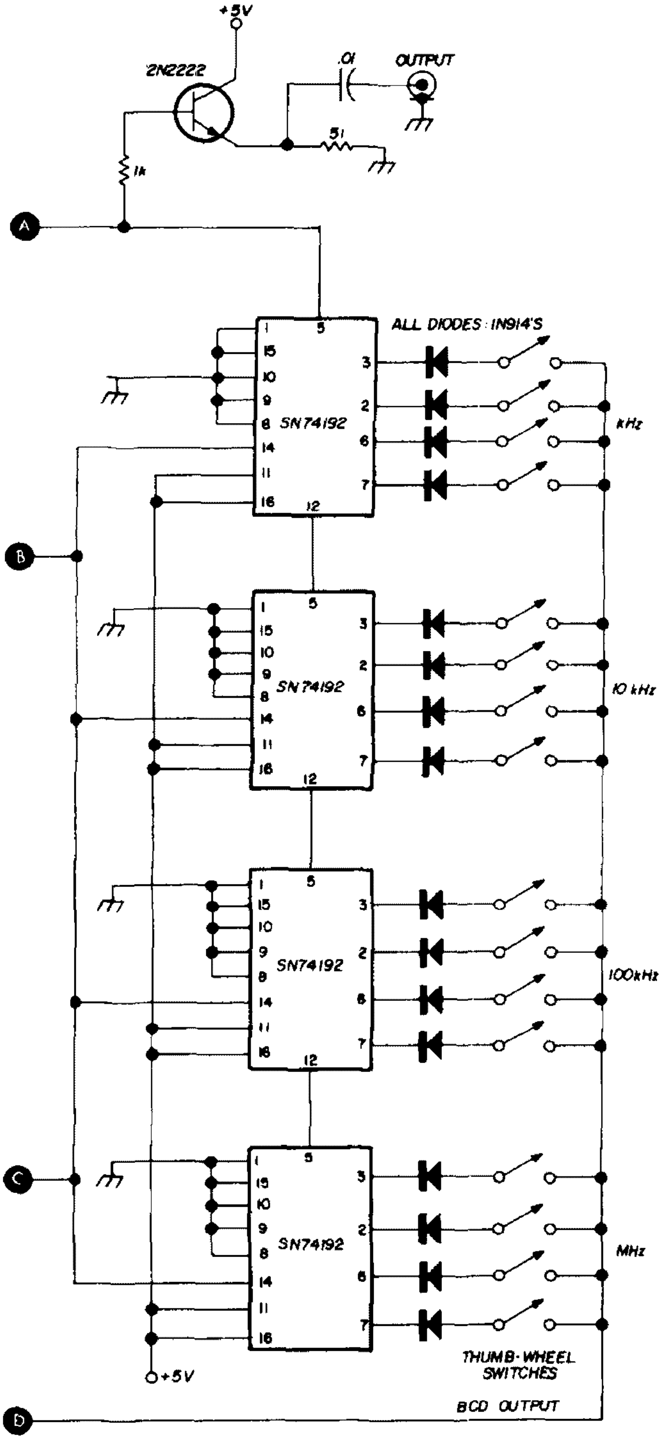


fig. 3. Schematic diagram of the phase-locked loop frequency synthesizer. A major parts list for this unit is given in table 1.

plex integrated digital and analog circuit. Its function will not be completely explained here, but a complete explanation may be found in Motorola Application Notes AN541 and AN535. The frequency discriminator functions as a frequency-to-voltage converter. If the input frequency is *below* the reference, the discriminator produces a *high* output voltage; if the input frequency is *above* the reference, the discriminator produces a *low* output voltage. This voltage is called the control voltage, and determines the vco operating frequency.



Refer to fig. 1 and consider the following: assume the vco is operating at 2.50 MHz, and the programmable divider is set at 3000, corresponding to a frequency of 3.000 MHz. The output frequency of the divider is

$$f_{out} = \frac{f_{in}}{N} = \frac{2.5 \times 10^6}{3.0 \times 10^3} = 833 \text{ Hz}$$

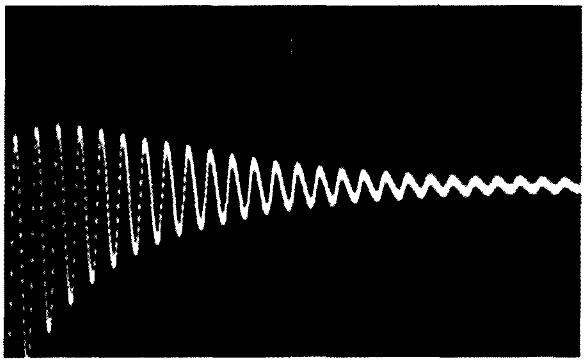
table 1. Parts list and cost breakdown for the frequency synthesizer.

qty	component	price	total
4	SN74192	$2.25	$9.00
3	SN7490	1.40	4.20
1	SN7400	.33	.33
1	MC4024	2.60	2.60
1	MC4044	2.60	2.60
1	SN74121	1.00	1.00
11	IC sockets	.40	4.40
2	2N2222	.50	1.00
1	filament transformer	1.49	1.49
1	LM309K	2.00	2.00
1	1-MHz crystal	4.50	4.50
4	thumb-wheel switches	2.25	9.00
Resistors, capacitors, cabinet end plates, etc.			10.00
			$54.12

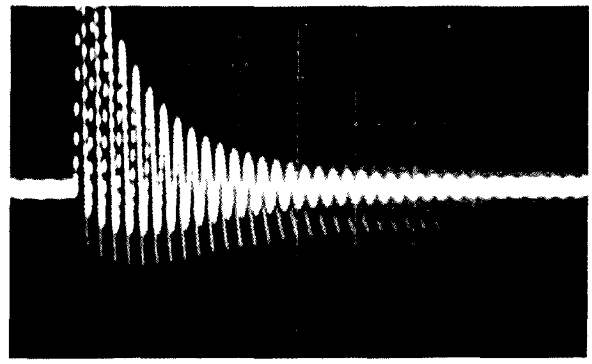
This frequency is lower than the 1000-Hz reference. Therefore, the frequency discriminator will produce a *high* control voltage which will raise the vco frequency. The vco will continue to change until the vco frequency is exactly 3.000 MHz. When that happens, the system is said to be phase-locked. If anything should happen to change the vco frequency the frequency discriminator will correct the fault.

A commercial frequency synthesizer is a very expensive piece of equipment. It must be assumed that such an inexpensive unit as the one described in this article must have shortcomings, and indeed it does. This unit has a range of about 300 kHz to more than 10 MHz in 1-kHz steps. A commercial unit may have a range of a few Hz to hundreds of MHz in fractional Hz steps.

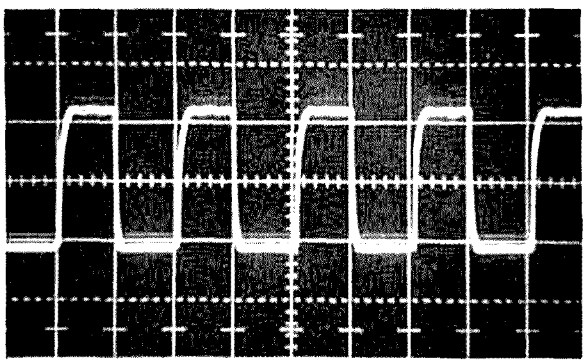
This degree of resolution and wide range is not necessary for amateur builders; neither is the corresponding price tag. The output of my unit is a square wave which may also be a disadvantage. The unit also has a certain amount of frequency modulation generated by noise voltages in the system. For *general* receiver and transmitter alignment



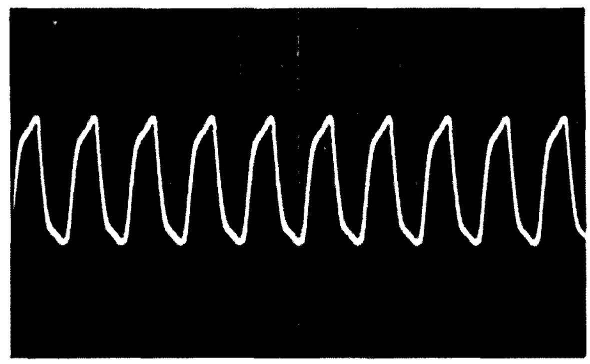
A. Vco control voltage with step frequency change from 510 kHz to 500 kHz.



B. Vco control voltage with step frequency change from 500 kHz to 510 kHz.



C. Output voltage at 500 kHz.



D. Output voltage at 5.000 MHz.

Waveforms of the frequency synthesizer. In A and B time base is 100 milliseconds per division; vertical scale is 50 mV per division. In C and D time base is 1 microsecond per division; vertical scale is 0.5 volt per division. Tektronix 545B oscilloscope with CA plug-in and X1 probe.

this fm is not objectionable. However, the synthesizer should not be used for direct transmitter control.

construction

The entire synthesizer is built on perforated fiber boards as separate modules, carefully shielded and filtered to reduce noise to a minimum. The reference oscillator and divider, the vco and the frequency discriminator should be mounted in separate aluminum boxes

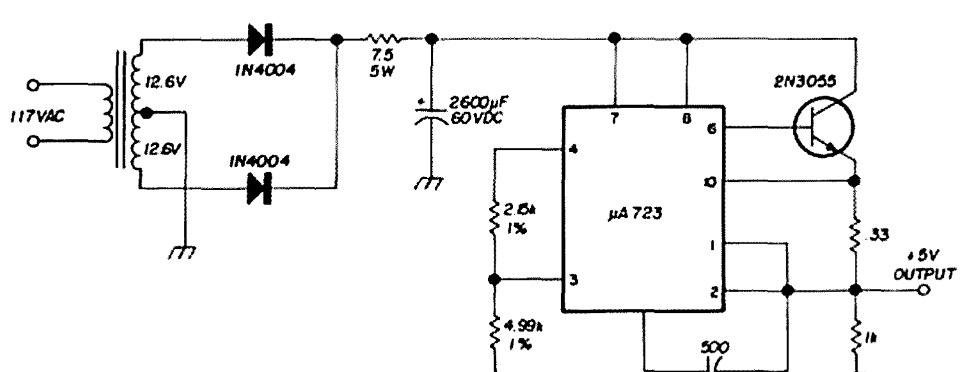
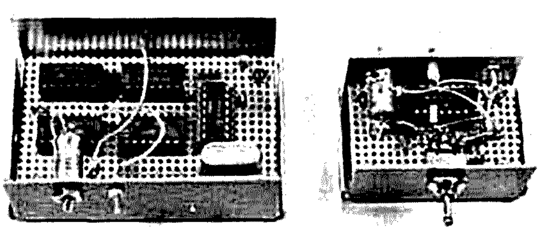


fig. 4. Power supply for the frequency synthesizer.

(see fig. 3). The power supply and programmable divider are mounted on the rear and front, respectively, of the cabinet.

It is absolutely necessary that all ground leads for the power supply and the individual modules be returned to a common point to discourage ground loops. This system is susceptible to even the slightest ac ripple, and ground *loops* are difficult to analyze.

Likewise, a topnotch power supply is in order. My synthesizer uses a μA723 voltage-regulator IC and a pass transistor, although the much easier to use LM309 was tried with success. Both devices are currently available at bargain prices.



The 1000-Hz reference source is on the left, and the voltage-controlled multivibrator is on the right.

If you want a sine wave output, it may be obtained by feeding the square wave into a tuned circuit resonating at the fundamental, or any one of the odd harmonics. However, most alignment jobs can be handled well with the square wave output. In fact, the harmonics are useful for extending the range of the unit. It should be borne in mind, however, that the fm noise is multiplied by the order of the harmonic, so that noise will be five times as bad at the fifth harmonic as at the fundamental, three times as bad at the third, and so on.

The synthesizer is easy to build and is indispensable as a tool for general receiver and transmitter alignment. It is well worth the $50.00 or so it costs to build.

ham radio

an efficient all-band tuned-dipole

Getting maximum efficiency from an outstanding multiband antenna

E.R. Cook, ZS6BT, 32 Grove Road, Gardens, Johannesburg, Republic of South Africa

The most popular antenna for 80 and 40 meters is the dipole. For the higher frequencies, the rotary beam is the favorite. However, there is much to be said for the all-band antenna, and the one described here will outperform a three-band Yagi in the dipole's favored direction and will only be about 6 dB down from the Yagi off the dipole's ends.

The average three-element three-band beam is actually very inefficient in transferring power from the transmitter into the ionosphere; it merely transmits its radiated power in a focused beam. Losses due to standing waves, mismatch and imbalances all reduce the efficiency so that the gain over an efficient reference dipole is not as great as it seems.

A well-balanced, highly-efficient tuned dipole, properly matched to the transmitter, radiates more power than does the untuned, unbalanced, coax fed dipole of a three-band beam. However, it wastes a lot of energy in unwanted directions while the less-efficient tribander makes the best use of whatever power it does radiate. Overall, the tribander does the better job.

The purpose of this article is to describe an all-band dipole which gives normal performance on 80 and 40, but competes favorably with a three-element beam on the higher frequencies. Apart from its ability to lay down a good signal, the antenna has attributes not possessed by a tribander.

basic design

The basic design is simple and well known. Take two pieces of 14 gauge wire 132 feet long, and use anything from 54 to 65 feet of each to form half of a center-fed dipole between 108- and 130-feet long. Use all of the remaining wire to build an open-wire transmission line with six-inch spacing. How to dispose of the feedline is a problem for the individual to solve. The result is an open-wire-fed dipole which is, effectively, a half wave on 80 meters with quarter-wave feeders. If your masts are less than 108 feet apart, arrange matters so that 8 or 10 feet at each end of the dipole drops vertically, but stop these dangling ends from swinging in the wind.

The feedline at ZS6BT is 78-feet long, and the antenna is 40-feet high. There is a six-foot length of feeder in the shack, and 30 feet goes vertically to the dipole. The remaining 42 feet runs horizontally at a height of 10 feet and is strung between the shack and a pole 10 feet high. It is held taut by a pair of turnbuckles, and there are no spreaders on the horizontal feedline.

The antenna is tuned by a balanced-to-unbalanced Z-match tuner connected through six feet of 50-ohm coax to the transmitter. Because of the quarter-wave feedline on 80, parallel tuning is used on all bands. There is no need to describe the tuner as many good designs have already appeared in the various amateur magazines and in the *ARRL Handbook*. The Johnson *Matchbox* does the job well and the tuneup information in this article is based on the *Matchbox*.

It is essential to use an swr bridge in the coax between the transmitter and the tuner. A twin-meter frequency-independant instrument is ideal, but any bridge will do. The idea is to tune the antenna to frequency anywhere in any band and to reduce the standing wave on the coax to zero. This produces a very efficient antenna.

There is no need for diagrams at this stage. A beginner could put up the antenna from the written description, and he would have an excellent all-band antenna. There are probably many such already in use. Radiator length is not important, as can be seen, and we need not be particular to 6 inches with the 132 feet of wire. The secret does lie in having quarter-wave feeders.

What we have done, to date, is describe a normal antenna which is an excellent performer on 80 and 40, and a reasonably good performer on all bands 80 thru 10. By strapping the feeders together and working against a good ground, it does an excellent job on 160 as a T. At ZS6BT, I copy North Americans on 160 and have been heard on that band at 2,000 miles with 10 W input.

improvements

Now, to that wasteful radiation from a dipole. There is nothing we can do about 80 and 40, but much we can do about the other bands. If we stack two dipoles, one above the other, and drive them in phase, we pull down a lot of that wasted upwards power and force it broadside to good effect. Stacked, driven dipoles and all-band performance do not go together very well.

If we sling a parasitic reflector, a half wavelength long and a half wavelength below a dipole, we have the proper phase for broadside performance, even though we do not have the efficiency of the driven array. Moreover, this reflector interferes with the ground effect and tends to alter the angle of radiation beneficially. Once we consider that we can sling two such reflectors in-line for 20, three in-line for 15 and five in-line for 10; we see a chance for considerable improvement.

Neither reflector length nor spacing is all that critical. In fact, if we narrow the spacing we may correct the phase by lengthening the reflector. If we have not sufficient head-room on 20, we may lift the reflectors another 5 feet and lengthen them by about 5 feet without ill effect.

Fig. 1 shows the arrangement and normal dimensions. The idea can be tried out on existing dipoles too.

tuneup

The initial tuneup of the antenna should be done with the aid of a dummy load, in order to calibrate the antenna tuner, and thereafter band changing will be a simple matter.

On each band, tune and load the transmitter into a dummy load until the swr bridge shows maximum forward current; of course, there should be no sign of reflected current.

Do not touch the *transmitter tuning* and loading controls, but couple up the antenna and tuner and vary the tuner controls until you obtain (a) the same forward current that was seen with dummy load and (b) no sign of reflected current. You see why a twin-meter bridge is best; you can read both directions simultaneously. Make a note of the tuner dial settings for future use.

In normal use, set the tuner controls appropriately and load up. Then tune out the last vestiges of reflected current and leave the bridge measuring reflected power. As you move around in the band, see that you maintain a zero on the bridge meter.

Now, about the bonuses which the tuned dipole can offer. It will attenuate harmonics because it is not tuned to them. If you use a TVI filter the extra tuned circuit between transmitter and antenna will improve the filter performance.

On reception, be sure to tune the antenna to the right band or you may get poor results; you have an extra tuned circuit between antenna and the receiver input. For some, this may be an extra bonus.

Modern receivers do not suffer greatly from images, but there is a possible exception. With shortwave broadcasts running to megawatts, they can break through even though the first i-f is 4 MHz or so. They will do this more easily if the antenna will respond to the broadcast station's frequency. Even a three-band beam is not very frequency conscious on reception, in part due to pickup on the coax outer shield, but a tuned dipole tends to attenuate signals to which it is not actually tuned.

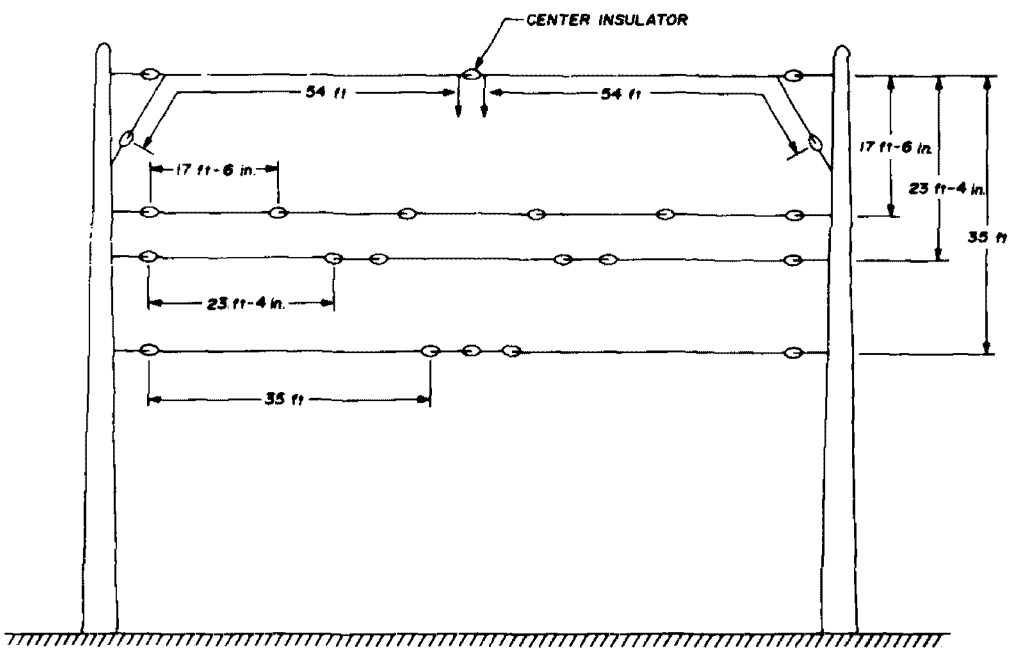


fig. 1. General arrangement and nominal dimensions for ZS6BT's all-band dipole.

There is no need to use particularly heavy guage wire for the reflectors as they are under no strain. The 20 meter, and perhaps the 15 meter reflector may be tied to the masts. The 10 meter ones may require halyards. To prevent unnecessary absorption, it is better to use nylon cord for guy wires and halyards.

160 meter operation

The first essential for 160 is a really good ground connection. Basically, the

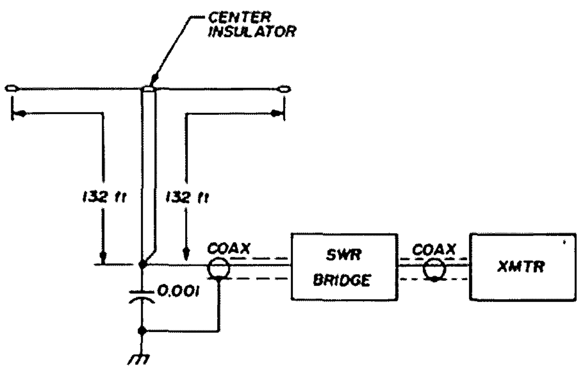
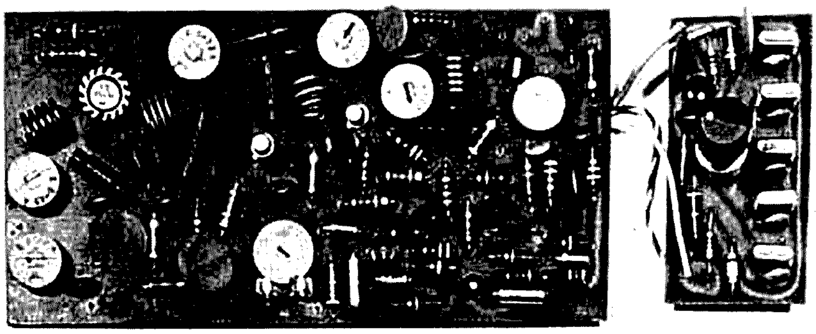


fig. 2. Arrangement for using the all-band dipole on 160 meters.

matching to the coax is done by an L-network. However, with a T antenna of the dimensions given, the L inductance appears to be unnecessary and only the capacitor is used at ZS6BT. Aim for maximum forward current and no reverse current on the bridge. My system is shown in fig. 2. The capacitor is an Aerovox Series 1650, 1200 V, 0.001 mF. The matching unit, therefore consists of a coax socket and the mica capacitor. Individual installations would call for some experimentation regarding matching.

During the past 45 years, most of the usual, and some unusual, antennas have been tried; most were up less than a year and not one gave six-band results which were acceptable. The present antenna has been in use for over eight years and there is no intention of trying any other. At last I am satisfied.

ham radio



five-frequency crystal deck for the Sonobaby

By popular demand, here is how to add frequency flexibility to the Sonobaby, and how to make a handy signal generator

Robert Shriner, WAØUZO, Robert Heptig, KØPHF, Earl Gill, WBØEDA

The Sonobaby transmitter, featured in the October 1971 issue of *ham radio,* was an all-transistorized two-meter fm transmitter. Since the article was published, we have had numerous requests for a similar multiple-frequency transmitter. This is understandable in areas served by many repeaters or by the mobiler wishing to operate through the different repeaters encountered in his travels.

The modification described here is applicable to the Sonobaby and other transmitters using 18-MHz crystals, and provides five-channel capability. No parts of the original transmitter need be thrown away and the five-frequency deck is small enough to fit into a hand-held portable as well as a base or mobile unit.

As an added feature, the five-frequency deck can be used as a five-frequency signal generator. In order to do this a separate audio board has been developed allowing the signal generator to have an unmodulated carrier, a constant tone or a microphone input.

The circuit was designed with the idea of not wasting any parts while still enabling the original Sonobabys to be used on five frequencies. The original Sonobaby oscillator is changed to a buffer stage and a new oscillator is built on the five-frequency deck. Mechanical switching offers design simplicity and economy. For those who want only two or three frequencies, the answer is simple – only mount the crystals and trimmer capacitors desired. If more frequencies are needed later, they can be added easily at that time.

The five-frequency deck is soldered directly to the switch tabs thus minimizing lead length capacitance.

Audio and operating voltages are taken from the Sonobaby. The rf output of the deck is taken from the collector of the new oscillator and fed via 50-ohm coax

*An etched and drilled circuit board for the five-frequency deck is available for $2.00. Boards for the frequency deck and audio stage, suitable for making a signal generator, are available for $3.50. Write to Sonobaby, Box 969, Pueblo, Colorado 81002.

and a capacitor to the base of the original Sonobaby oscillator.

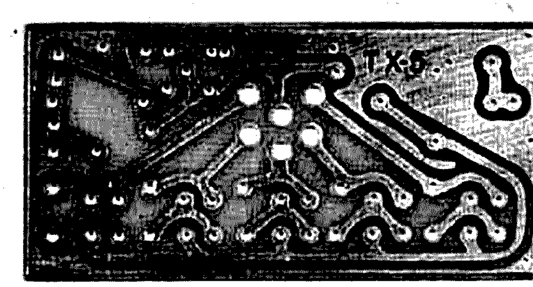
An etched and drilled circuit board is available for the five-frequency deck, or you can make your own following the layouts in the photos.* Construction is very easy and should not take more than an hour. As in all printed-circuit work, clean the board thoroughly first. An ordinary pencil eraser rubbed over the board will facilitate soldering and is well worth the few minutes it takes.

To get started on the five-frequency deck, use a pair of long nose pliers to break off the tab on the back of switch S1 that goes to terminal 6. Mount the switch on the circuit board with the tabs protruding through *to the foil side.* Solder the tabs of the switch to the circuit board.

Install the remainder of the components, with the exception of trimmer capacitors C7 through C11, following the photographs and schematic layout. Be very careful about applying too much solder which might tend to bridge the circuitry. Also, clip all the leads as short as possible after soldering. After a thorough examination of the board for bridged circuits and bad solder joints, install trimmer capacitors C7 through C11. Note that the trimmer capacitors are installed on the foil side of the circuit board instead of the epoxy side as are all the rest of the parts.

Refer to the October 1971 issue of *ham radio* and the schematic diagram of the Sonobaby transmitter. Remove CR1, Y1 and C12. Note that these parts are used on the five-frequency deck, so don't throw them away.

We recommend that at this time you mount the five-frequency deck on your chassis with the original Sonobaby. Mounting is done by drilling a 1/4-inch hole in your chassis in a convenient location. Make sure there is enough room inside the chassis for the circuit board.



Foil side of the frequency deck board. Parts layout can be gleaned from the other photos.

Audio is taken from the Sonobaby at the junction of L1 and C8. Run it over to the audio input of the frequency deck with a piece of shielded wire. Take the operating voltage from the Sonobaby at the junction of R15 and C12. A short piece of hookup wire will do.

The rf output from the oscillator on the frequency deck goes to the point where the crystal Y1 was originally connected to the junction of C9 and C10.

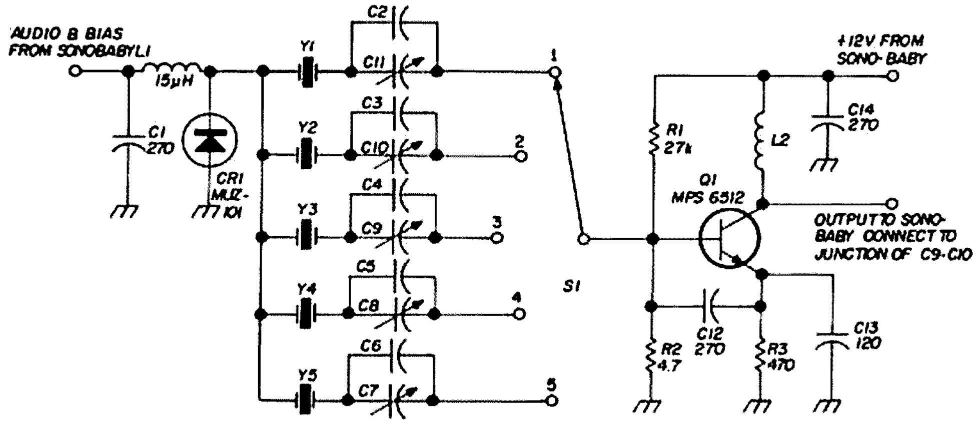


fig. 1. Schematic of the five-frequency deck for the Sonobaby.

C2-C6	15 pF NPO	S1	1P6T switch, Daven 18-KM
C7-C11	5.5 to 18 pF, Erie 538-011A-5.5-18	Y1-Y5	see text
L1,L2	15-μH choke, J.W. Miller 9310-40		

Use a piece of 50-ohm coax for this run and keep it as short as possible – not over ten inches.

Adjust C9 on the original Sonobaby for maximum capacitance, check everything carefully and you are ready to go. Capacitors C2 through C6 are used to put

We ordered our crystals from International Crystal. Besides specifying operating frequency and commercial standard tolerance, mention that the crystals are for a Westinghouse Air Brake Company Carry-phone II, 20TS-1 transmitter.

If you want to go hog wild and have

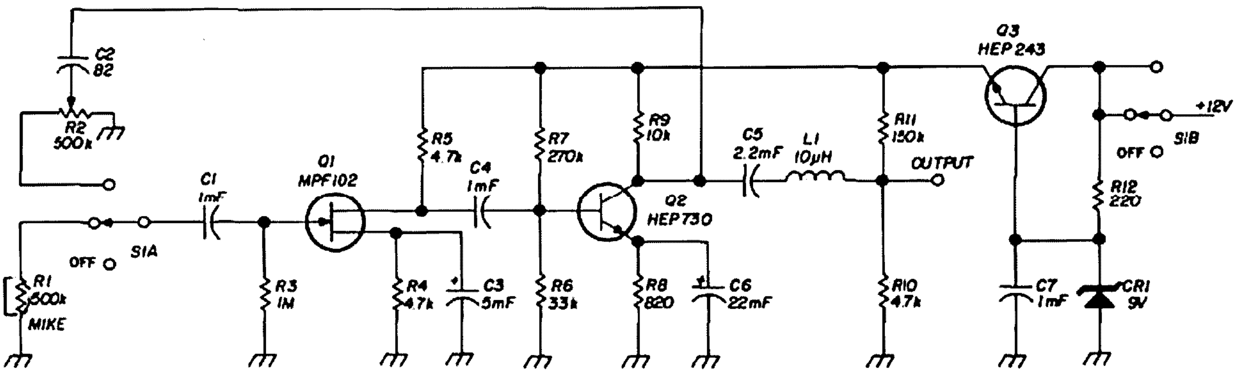
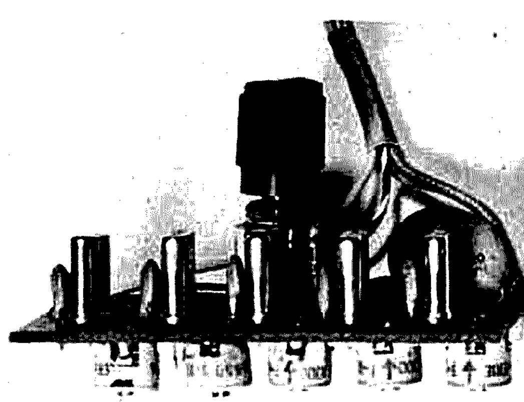


fig. 2. Schematic of the audio board used to turn the frequency deck into a versatile signal generator. C2 can be varied to suit the feedback frequency desired. R1 and R2 are deviation controls. Using quarter-watt, 10% resistors for R3 through R12 helps keep things compact.

each crystal on frequency. Adjust C9 on the original Sonobaby board for best injection voltage to the buffer stage. Check for spurious radiation and adjust C9 for maximum power output and minimum spurii.

We have had no difficulty with the installation of these five-frequency decks. It is a good idea not to attempt to locate the deck over a foot away from the Sonobaby transmitter – the closer the better. The tuning of the Sonobaby should be touched up and it should be peaked on one of the lower frequencies to be used.



Side view of the frequency deck shows switch mounting and the netting capacitors.

more than five frequencies, merely install two decks and you'll have ten frequencies. In this instance, set one of the decks at switch position 6. This will turn it off and allow you to use the other board for the next five frequencies.

signal generator

To construct a signal generator that will cover the two-meter band, merely take the frequency deck and add an audio board, fig. 2 This board supplies the regulated voltage required by the varactor CR1 on the frequency deck. S1A on the audio board selects either a steady tone to the signal generator or a microphone input. A crystal or hi-impedance mike will work fine.

About 50 microvolts of signal can be taken from C1 of the frequency deck, and as the signal is rich in harmonics, it can be injected into the front end of a receiver for tuneup and alignment.

We would like to take this opportunity to thank all of you for your letters praising the original Sonobaby. It is very gratifying to know that we have helped so many get on the air and that you have been so well pleased with the circuit.

ham radio

pulse snap diode impulse generator

Basic data and practical applications of a new and versatile addition to the solid-state diode family

Bernard S. Siegal and Michael K. Turner, Siliconix, Inc.

One of the many interesting devices to make the scene in the world of solid-state components is the pulse snap diode. Also known as a step-recovery diode, this device was included in Hank Olson's review of frequency multipliers that appeared in an earlier issue of *ham radio.*[1] The diode described in the article was the Siliconix SV110. Its use is by no means limited to the application shown, and it's priced within the means of most experimenters. Let's take a look at the SV110 and see what makes it snap.

features

The SV110 was designed to meet the need for an impulse generator that would produce a high-amplitude pulse (up to 70 volts) of subnanosecond width. Some of the applications of the SV110 include:

1. An rf calibrator.

2. Local-oscillator frequency source.

*Since the time this article was submitted, Siliconix has dropped the pulse-snap diode product line. Suitable replacements for the SV110 are the Microwave Associates Snap-varactor 4756, and the Hewlett-Packard Associates Step Recovery Diode 5082-0180. Some slight variation in circuit element values may be required to accomodate either of the two suggested replacement diodes.

3. Accurate signal source for testing components.

4. Clock generator.

5. Phase-locked-loop signal source.

6. Sampling-gate pulse source.

In short, the SV110 is the answer for the designer looking for a stable signal source with usable harmonics over a wide frequency range.

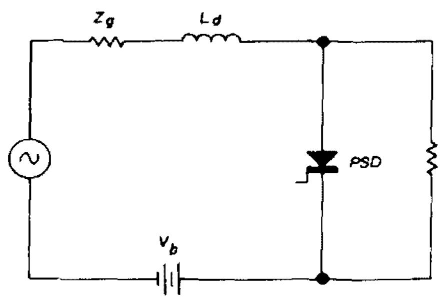


fig. 1. Basic circuit using the pulse snap diode.

basic circuit

All applications of the pulse snap diode rely on the charge-storage characteristics of the device. During forward conduction, the diode stores charge and exhibits a very low impedance (typically less than one ohm). The device maintains its low-impedance state with reverse bias until the charge stored in the junction is depleted. At the instant charge is depleted, the diode will switch from its low impedance state to a high-impedance state.

The time required to complete the change in impedance (approximately 300 picoseconds) is called the transition time. This parameter depends on the amount of charge stored in the diode. More details on operational characteristics will be found in references 2 through 5.

A basic circuit of the impulse generator is shown in **fig. 1**. Parameters are defined as follows: Z_g is the generator impedance, L_d is the driving inductance, Z_l is the load impedance, and V_b is the bias voltage. **Fig. 2** shows waveforms at various operating points in the circuit.

The generator voltage, e_g, and current, i_g, waveforms appear in **fig. 2A** and **fig. 2B**. Generator current i_g lags e_g by approximately 90 degrees due to the inductance, L_d. When V_b is zero, the diode current, i_d, waveform will be the same as the i_g waveform in **fig. 2B**. Charge is stored in the diode junction during the positive half-cycle and removed during the negative half-cycle. The amount of charge removed must always be equal to the charge stored, less recombination losses.

When V_b is greater than zero, as in **fig. 2C**, the diode will not begin to store charge until the voltage across the diode is greater than the sum of V_b and one diode voltage drop. For any given input, the greater the value of V_b, the less charge is stored in the diode, as the diode is forward biased for increasingly smaller portions of the positive half-cycle.

impedance transition

For values of V_b greater than zero and less than the maximum value of e_g minus one diode drop, the time required to remove the charge stored in the diode will

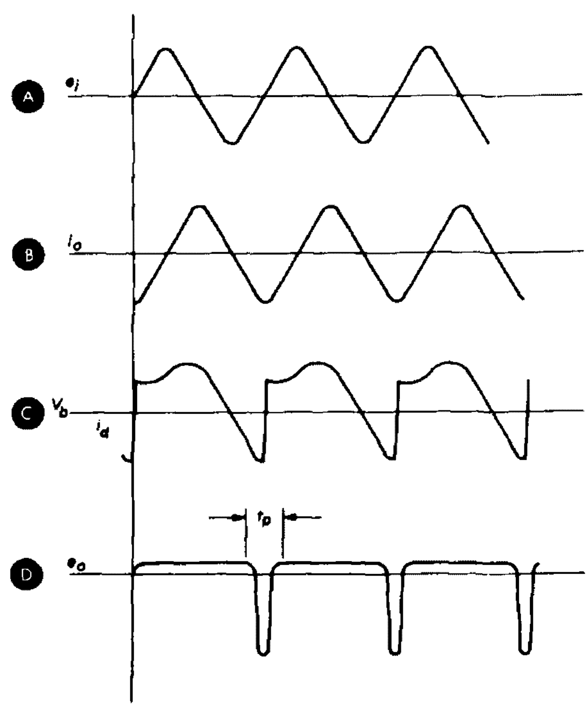


fig. 2. Waveforms at various operating points in the circuit.

be less than one half-cycle. The diode will return to its high-impedance state even though i_d is a negative, non-zero value. By proper selection of V_b, the transition can occur when i_d is at its maximum negative value as shown in fig. **2C**; this is the point of maximum pulse energy.

A 50-ohm load was used for test equipment convenience. Any value of Z_L between 25 and 75 ohms could be used with only total power output affected. Design equations for Z_L can be found in Hewlett-Packard Application Note 920[3]

At transition, the energy stored in L_d will charge the reverse capacitance of the diode, and a ringing waveform will appear across the load (fig. 2D). The characteristics of the ringing waveform are determined by the resonance of L_d and the reverse capacitance of the diode, in combination with the load impedance. Only a single negative pulse is generated for every input cycle, since the positive excursion of the ringing waveform is clipped by the diode. A time-domain profile of the output would show a comb spectrum with a spike at every harmonic of the input frequency. The spectral envelope will show a decay in amplitude until a null is reached at 3/2 t_p, where t_p is the pulse width.

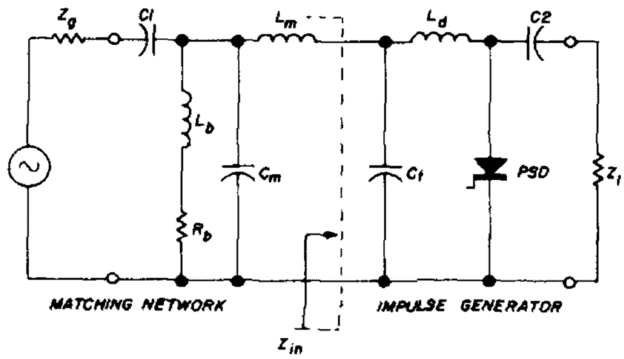


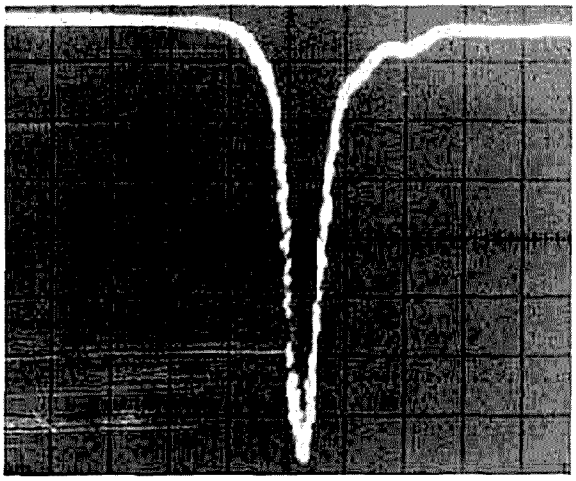
fig. 3. Practical circuit of an impulse generator using the SV110. Component values are described in the text and table 1.

practical circuit

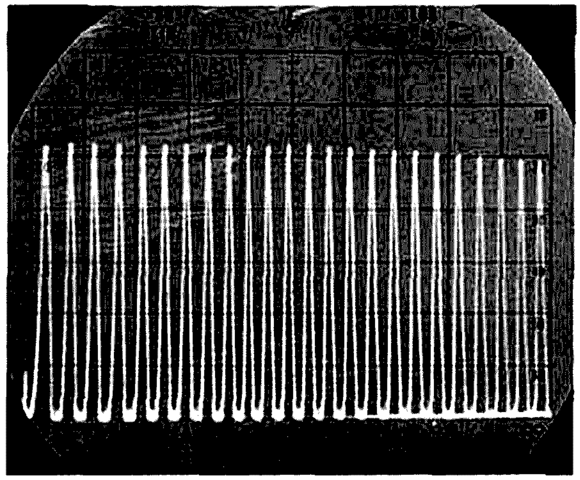
A schematic of a practical impulse generator is shown in fig. 3. Parameters C_t, L_d, and the pulse snap diode, PSD, form the impulse generator. The driving inductance, L_d, resonates with the reverse capacitance of the diode to obtain the output pulse. L_d should include any stray

table 1. Calculated values for components in fig. 3.

parameter	equation	example	remarks
t_p	see text	2ns	Determined by application
L_d	$L_d \approx \frac{(t_p{}^2)}{\pi} \cdot \frac{1}{C_{TR}}$	100 nH	C_{TR} = reverse capacitance of PSD = 4 pF
C_t	$C_t = \frac{C_{TR}}{(2f_{in}t_p)^2}$	2500 pF	
R_{in}	$R_{in} = 2\ f_{in}L_d$	6.28Ω	Required for matching network
L_m	$L_m = \frac{Z_g Z_{in}}{2\pi f_{in}}$	283 nH	
C_m	$C_m = \frac{1}{2\pi f_{in} Z_g Z_{in}}$	900 pF	
R_b R_b	$R_b =$ $R_b = \frac{2\tau}{\pi N^2 C_{TR}}$	25Ω	τ = minority carrier life-lifetime (ref. 6) $N = \frac{1}{2f_{in}}/t_p$



Pulse profile from the circuit of fig. 3. Pulse width is near the calculated value of 2 ns.



A slight variation in pulse power is shown, which is caused by uncompensated stray inductance in the test circuit.

inductance as well as the diode package inductance (reference 6). C_t resonates with L_d at f_{in}, the input frequency, for maximum power transfer. C_t must present a good rf short circuit to the output frequencies.

L_m, C_m comprise a low-pass matching network between the input impedance and Z_{in}, the impedance of the pulse generator. The matching network is bandwidth limited:

$$BW \cong \frac{2f_{in}}{(Z_g Z_{in} + 1)^{1/2}}$$

where BW is the bandwidth. Wideband matching networks may be substituted[7] to obtain a wider spectrum of input frequencies.

bias network

It is important that series resonances are not introduced into the bias network, L_b, R_b (fig. 3). Such resonances, especially those below f_{in}, may cause the diode to act as a negative resistance, which would probably result in spurious signals in the output. L_b should present a high impedance at f_{in}. A commercially available molded choke with a self resonance at f_{in} will give the greatest isolation between the bias network and the input to the impulse generator.

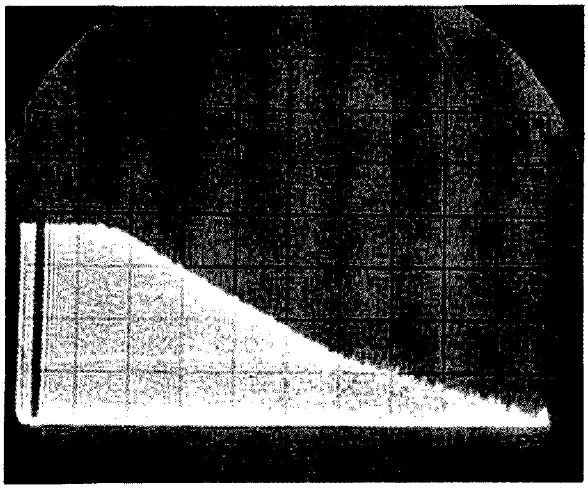
R_b is shown as a simple resistor; however, R_b could be replaced with a variable resistor, external bias-voltage source, or a combination of external and self bias.

impulse width

The impulse width, t_p, is important. All critical component values, as well as power variations between harmonics in the impulse train, are determined by t_p. The following design example shows the importance of t_p.

A frequency of 10 MHz was chosen for f_{in}. Any frequency could have been selected, providing the period of f_{in} is less than the minority-carrier lifetime of the PSD (*see reference 6*). For the SV110, the minimum value of f_{in} is approximately 5 MHz. Power variations between f_I and f_{20}; i. e., between the input and its 20th harmonic, were held within 1 dB. A flat power response is desirable if the impulse generator is to be used as a signal source for calibration work. To provide this small power variation over a large number of harmonics, t_p must be quite small. Reference 3 provides a graphical method of finding the required t_p.

For the example mentioned, t_p was approximately 2 nanoseconds. Values for minority carrier lifetime, τ =200 ns, and reverse capacitance, C_{TR} = 4 pF, were obtained from the SV110 data sheet.[6] With this information and that in table 1, the component values for fig. 3 were obtained.



The entire spectral envelope, or comb line. Discontinuity near beginning of envelope marks start of comb line.

test results

The circuit of fig. 3 was constructed on double-clad copper board. Stray capacitances and inductances were kept to a minimum. The oscillographs show circuit performance. Pulse width is very near its calculated value. Variations from the calculated value are caused by uncompensated stray inductance. Power variations in the harmonics can be traced to the same cause.

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5. R. E. Winkelman and E. G. Cristal, "Filter Matches to SRD Impulser," *Microwaves,* September, 1970, pp. 34-38.
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ham radio

adding 160-meters

to a 40-meter vertical

Making an inexpensive multi-band antenna with excellent 160-meter performance and hidden benefits

Ken Cornell, W2IMB, 332 West Dudley Avenue, Westfield, New Jersey 07090

It is a myth that only lighthouse keepers *and cattle ranchers can get on 160 meters* because of the real estate needed to erect a really effective 160 meter antenna. *Anyone* can get up an inexpensive and effective 160-meter antenna in a limited space. For my first operating on 160, I used my 80-meter doublet with the feed end shorted and worked against ground as a T antenna. An improvement over this was an end-fed L run out my basement window, up the side of my house and across my yard. It also was operated against ground — provided by a convenient cold water pipe in my basement station.

outline

As I started working real DX with this last scheme — 8's and 9's — I noticed that the best signals on the band invariably were using verticals. I decided to try adding 160 capability to my reliable 40-meter vertical standing in my back yard. The basic plan was to add a 40-meter trap at the top of the existing vertical, add a piece of plastic pipe wound with a 160-meter loading coil and top this combination off with an 8-foot section of aluminum tubing as a top load. I could drive a ten-foot ground stake and use the antenna for both 40 and 160.

40-meter vertical

My 40-meter vertical consists of some World War Two surplus mast sections, labeled AB-85/GRA-4. Each section is 36-inches long, 1 5/8-inch diameter and has a wall thickness of 1/8-inch. One end of each section is swedged down to 1¼-inch diameter for about a six-inch length where the sections join together. The female ends of each section also have four wiper springs that help to insure electrical continuity. These were selling quite inexpensively and I picked up several hundred feet of them one Field Day. While these sections might be hard to find, almost any type of aluminum or galvanized steel pipe will work for the lower 32½-foot mast section. I do not recommend aluminum tubing with less than 1/8-inch thick walls (12 gauge). Most of the aluminum TV mast sections I have seen look pretty skimpy, but could probably be used with proper guying and support.

construction

The first order of business was to find the plastic pipe for the loading coil. I called several warehouses that stock this item, but they have a $25 minimum order gimmick. I finally found a local plumbing supply house that had one piece of 1¼-inch PVC pipe twenty-feet long. I had to take the whole piece (for $7.80), but they cut it into three equal pieces for me. The 1¼-inch inner diameter of this plastic pipe made a nice tight telescopic fit to the swedged end of my mast section.

If you have trouble getting the proper inside and outside diameters of the plastic pipe and tubing, I would suggest that you get the plastic pipe with a larger inside diameter than your top and bottom mast sections. By wrapping several bands of tire tape spaced about 4- to 6-inches apart, you can adjust the diameter of the masts for a snug telescopic fit into the plastic pipe. Once fitted, secure the joints with a bolt.

Assembling the mast material should not be difficult. Remember to install a bolt and solder lug at the top of the 32½-foot mast, one about three inches above the bottom of the plastic pipe (jumper these two lugs), one about three-inches higher (for the 40-meter trap), one at the top of the plastic pipe and one at the start of the 160-meter loading coil (jumper these last two lugs).

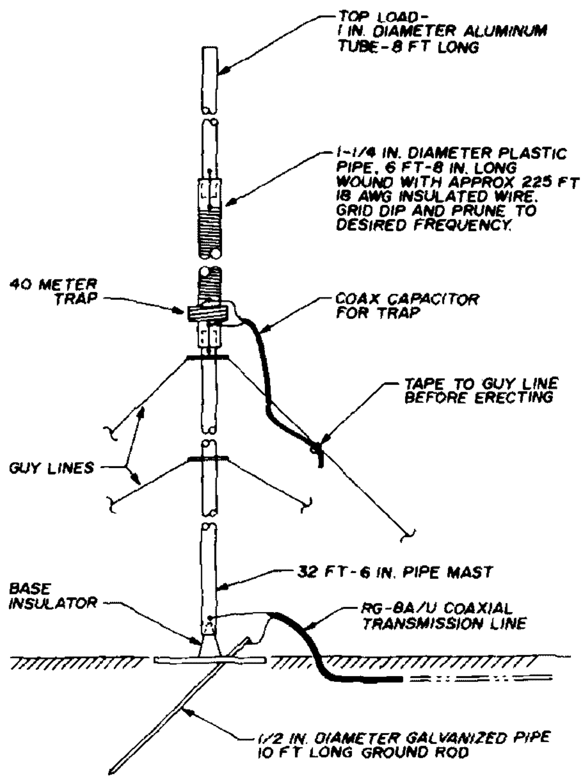

fig. 1. The arrangement of the 40- and 160-meter antenna.

The lugs and jumpers provide the electrical continuity between the masts and the coils.

40-meter trap

I had a few 40-meter traps left over from an old Field Day antenna. The traps were originally described in an old *QST* article and they used sections from commercially available air-wound coil stock. The coil in one trap consists of nine turns of number twelve wire with a 2½-inch diameter and with the turns are spaced

about 1/8-inch apart. The coil was shunted with a 100-pF high-voltage type TV capacitor.

Feeling that these capacitors might not weather well, I decided to use a piece of RG-8/U as the capacitor. RG-8 has a capacitance of 29.5-pF per foot and RG-II has a capacitance of 20.5-pF per foot. I cut a piece of RG-8 about four-feet long, dressed one end and soldered the shield to one end of the coil and the inner conductor to the other end of the coil. The grid-dip meter indicated resonance too low, so I trimmed the coax until I zeroed in on 7250 kHz. Be sure that the shield does not short to the inner conductor while you are trimming the cable. When you are finished, seal both ends of the coax with plastic cement to waterproof the cable.

Mount the trap concentric with the plastic pipe between the two lugs spaced at three inches, and just below where the 160-meter coil will go. Before erecting the antenna, tape the coax to one of the guy wires — allowing a little slack. Do not make the mistake I made of using RG-58 or RG-59 for the trap capacitor. It will work well with low power, but mine simply went *poof* when I fired up my linear on forty meters.

160-meter loading coil

Now for the fun of winding the 160-meter loading coil. *The ARRL Handbook* states that a helically-wound vertical antenna needs twice as much wire length as a normal quarter wavelength. Since my bottom section and top section added up to about forty feet, I figured I would need some 167 feet of wire in the trap to hit 1812 kHz. I had a new 175-foot roll of plastic-coated number-18 wire. I decided to wrap up the whole roll on the PVC pipe and grid dip it out to see where it resonated. The coil was more or less scramble wound and both ends were soldered to the lugs.

The next step was to check the assembly with a grid-dip meter. I leaned the antenna up against my garage, and using a step ladder, checked the coil for resonance. I was quite surprised that it tuned high. I then added another fifty feet of wire to the coil, and now I was too low. I began to prune ten turns at a time, then five and finally one turn at a time until I zeroed in on the desired frequency. This step requires patience, but for a good antenna, it is the most critical operation of all.

installation

The antenna is not too difficult to raise with a few helpers. The base, of course, should sit on a suitable insulator and foundation. I use two sets of non-conducting guys. One set is connected to a guy ring close to the top of the bottom mast section just below the PVC pipe and another set is connected half-way down the mast. I already had three 34-foot radials buried in the ground for the 40-meter vertical, but I added the ten-foot long half-inch diameter pipe as an extra ground for 160. We have a very heavy shale strata here about six feet below the surface, so I flattened one end of the ground rod and ground it sharp and drove it in at a 45-degree angle to get better exposure. I feed the antenna in the normal manner with RG-8/U buried under the turf.

operation

For some reason, this antenna works very well on all other amateur bands, including two meters. I also find that contrary to my expectation, it picks up less man-made noise than the inverted L, which is basically horizontal. I put this antenna up in the fall of 1970 and it has done a terrific job for me ever since. My basic feelings regarding the vertical versus the inverted L are as follows: Locally, the inverted L is stronger by several S points. From 400 to 800 miles I get conflicting reports. Over 1000 miles, however, the vertical vastly out-performs the horizontal. A final sneaky report, would be to state that the inverted L gives my wife some BCI, but I can be on the air all night with the vertical without a bit of BCI.

ham radio

new system for predicting six-meter sporadic-E openings

Here's a unique system for predicting six-meter sporadic-E propagation well in advance

Morrie S. Goldman, WA9RAQ, 5815 North Christiana, Chicago, Illinois 60659

The concept of using beacon transmitters to alert operators of the presence of vhf openings is certainly not a new one. For years, vhf enthusiasts have left their receivers parked on beacon frequencies, hoping to hear one of the relatively few low-power vhf beacons. Well known vhf experimenter Bob Cooper described such a beacon in *ham radio.*[1]

The use of amateur beacon transmitters is not, however, without its drawbacks. It goes without saying that one of the greatest problems with this early warning system is the low number of amateur beacons. It takes a devoted vhfer to build and maintain a device which will help others far more than it helps him.

Another reality is that it really isn't an early warning system. A warning yes — early, no. By the time you start to hear six-meter activity, the band is already open; not about to open! This means that you have probably already missed a great deal of the opening. Skip propagation is often quite selective, and the opening may have been to a relatively small geographic area.

If the notion of actually predicting a sporadic-E opening sounds far fetched, you're only partially correct. Long-range predictions for non-auroral related sporadic-E are still experimental, very involved and largely unproven. However, very short range predictions *are* practical. Predictions on the order of hours or minutes can be made with a high degree of accuracy and often you can pinpoint the affected geographic areas.

The sporadic E-(Es) maximum usable frequency (muf) begins low, and as ionization density increases, the muf increases. Es will appear on 11 meters before it developes on 10, and on 10 meters before reaching 6, and so on. However, just because it occurs on 27 and 28 MHz does not always mean that it will reach 50 MHz. During the summer months, you can hear a residual level of short-skip on Citizen Band — a mass of hetrodynes and squeals, and a multitude of accents from different transmitters. Opening on 6 meters, or for that matter, 10 meters, are not nearly so common.

So the question, and the main point of this article, is, how can we effectively put to use this known fact to predict Es?

Clearly, we need a more accurate means of following the maximum usable frequency. *It should also require little attention and be continuously operative.*

I began by discussing amateur beacons and their shortcomings. There is a substitute for these beacons which can be far more effective. It consists of transmitters staggered across the 30- to 50-MHz range, operating 24 hours a day with, by amateur standards, moderate to high power. These pseudo-beacons are provided by commercial paging stations.

These paging stations are operated by commercial companies to relay messages to their customers. Customers carry small receivers which are activated by a tone squelch or selective calling units, alerting them to repair orders, telephoned messages, etc. When there are no messages to transmit, many of the stations continue to transmit their call letters and location.

Vhf operators will be concerned primarily with those paging stations operating on 35.22, 35.58, 43.22 and 43.58 MHz. These four channels are the standardized low-band paging channels and will prove most effective for unattended monitoring.

Fortunately, receivers for this range are plentiful. They range from small inexpensive handheld portables to higher priced automatic-scanning monitors. Many older tube-type 30- to 50-MHz monitors can be found at bargain prices at hamfests. In many cases they are not as sensitive as the new portables, but most have effective squelch circuits and far better image rejection. These receivers are all designed for fm but a few have a-m-fm switches.

Most of the paging stations use a-m, which would seem to be a problem, but it isn't. All of the public-service-band receivers designed for consumer use that I've used have had such poor discriminator circuits that they also received a-m quite well. A well designed receiver might cause a problem, but I haven't run across one!

Unfortunately, all paging stations do not identify. Some operate on a tone-only system, and you never hear a human voice. Hearing a station of this type will alert you to Es, but not to where it's coming from. Some stations give their call letters in modulated CW but not their location. However, most pagers are still using voice and give both call and location. Many use names such as, "Paging Orlando," Air Call Chicago," "Fresno Radio Page," etc. After awhile you will start to recognize the voices of operators from some of the stations most commonly received in your area.

From various sources, including my own observations and those of members of *The Worldwide TV-FM DX Association*,* I have compiled a list of paging stations shown in table 1. It is not a complete list by any means, but it should give you a good start. Accurate lists are nearly impossible to obtain, and I believe that the latest record of paging stations kept by the Chicago FCC Field Office is dated 1963!

Callsign prefixes are not as clearly defined as in the amateur service as far as geographical area is concerned, but there is a general pattern, as shown in table 2.

The messages, callsign and other information broadcast by paging stations are usually recorded on a magnetic metal strip or endless tape cartridge, and continue to repeat until new messages come in. Many stations, as we mentioned earlier, transmit a recording of their callsign when there is no other traffic. Most of the MCW stations (or perhaps all) are off the air more often than on. When KSC645 in Chicago switched to tone-only operation, I finally began to hear weak signals on 35.58 MHz. Previously, only the strongest signals overrode the KSC645 signal. When the tones are on, however, nothing can make it through!

The procedure for using paging stations to monitor vhf conditions is a simple one. If you're using a tunable receiver, or a manually switchable crystal controlled receiver, set it to 35.22 MHz

*The Worldwide TV-FM DX Association publishes a monthly "Vhf-Uhf Digest" which includes technical articles, FCC news and reports on all phases of vhf and uhf DX. Sample copy is $.50; one-year subscription is $6.00 from WTFDA, Post Office Box 163, Deerfield, Illinios 60015.

table 1. Partial list of vhf paging stations in the United States. Corrections are welcomed by the author.

35.22 MHz

KIY757 a-m Birmingham, AL
KCH280 a-m Phoenix, AZ
KKX708 a-m Little Rock, AR
KMD342 a-m Fresno, CA
KMD998 a-m Lodi, CA
KMD681 a-m San Diego, CA
KMD305 a-m San Francisco, CA
KDN407 MCW Colorado Springs, CO
KCI299 a-m New Haven, CT
KIN645 a-m Miami, FL
KIY508 a-m Orlando, FL
KIY719 fm Pensacola, FL
KOK344 a-m Boise, ID
KSA623 a-m Ft. Wayne, IN
KSD320 a-m South Bend, IN
KAI934 a-m Des Monies, IA
KLB760 a-m Baton Rouge, LA
KKT407 MCW New Orleans, LA
KLB319 a-m Shreveport, LA
KGA807 a-m Baltimore, MD
KQD303 a-m Detroit, MI
KAH661 a-m Minneapolis, MN
KBM313 a-m Omaha, NE
KCC482 a-m Concord, NH
KEC925 MCW Buffalo, NY
KEA860 a-m New York City, NY
KIM905 a-m Charlotte, NC
KQD600 MCW Mansfield, OH
KGC266 a-m Allentown, PA
KGC223 a-m Philadelphia, PA
KFL880 a-m Greenville, SC
KKJ460 a-m Dallas, TX
KCF341 a-m Salt Lake City, UT
KCP253 fm Seattle, WA
KON908 a-m Cheyenne, WY
WWA335 a-m San Juan, PR

35.58 MHz

KOF328 a-m Tucson, AZ
KMD344 a-m Long Beach, CA
KME437 a-m Santa Cruz, CA
KMD347 a-m Stockton, CA
KAQ606 MCW Denver, CO
KIF651 a-m Ft. Lauderdale, FL
KIQ518 a-m Jacksonville, FL
KIE953 a-m Atlanta, GA
KIG844 a-m Augusta, GA
KUA217 a-m Honolulu, HI
KSC645 MCW Chicago, IL
KSC864 a-m Peoria, IL
KSD322 a-m Evansville, IN
KSD326 MCW Indianapolis, IN
KAD927 a-m Wichita, KS
KGA807 a-m Baltimore, MD
KCC266 a-m Springfield, MA
KQD601 a-m Flint, MI
KAD931 a-m Kansas City, KS
KEC935 a-m/MCW Newark, NJ
KED352 a-m Trenton, NJ
KEC519 a-m Rochester, NY
KEC515 a-m Troy, NY
KKT403 a-m Albuquerque, NM
KIY775 a-m Greensboro, NC
KKM248 a-m Oklahoma City, OK
KOA796 a-m Portland, OR
KGH861 MCW Chester, PA
KGC400 a-m Scranton, PA
KIG837 a-m Nashville, TN
KLB716 a-m Abilene, TX
KKV688 a-m Amarillo, TX
KKI445 a-m Houston, TX
KKQ965 MCW Lubbock, TX
KLB578 a-m San Angelo, TX
KIG297 a-m Norfolk, VA
KSD318 a-m Madison, WI

43.22 MHz

KMB309 a-m Los Angeles, CA
KMM960 a-m San Rafael, CA
KMM660 a-m Taylor Mountain, CA
KCC802 fm Waterbury, CT
KIN645 fm Miami, FL
KCI295 fm Manchester, NH
KEC745 MCW New York City, NY
KIY792 a-m Winston Salem, NC
KGC223 a-m Philadelphia, PA
KOP258 a-m Tacoma, WA

43.58 MHz

KOE257 a-m Phoenix, AZ
KMD986 a-m Sacramento, CA
KGA806 a-m Washington, DC
KIE367 a-m Miami, FL
KIG300 a-m Atlanta, GA
KSC644 a-m Chicago, IL
KSJ816 a-m Ft. Wayne, IN
KIF656 a-m Louisville, KY
KCB890 fm Boston, MA
KQC884 a-m Detroit, MI
KAF245 a-m Kansas City, MO
KAA893 a-m St. Louis, MO
KEA777 a-m Buffalo, NY
KEA627 a-m New York City, NY
KEC516 a-m Lafayette, NY
KEC518 a-m Rochester, NY
KQC877 a-m Cincinnati, OH
KQK593 a-m Cleveland, OH
KFJ891 MCW Columbus, OH
KGA804 a-m Philadelphia, PA
KGA805 a-m Pittsburg, PA
KIF653 a-m Memphis, TN
KKJ460 a-m Dallas, TX

table 2. Paging station callsigns by general geographic area.

KA	Midwest, including Colorado, Iowa, Kansas, Missouri, Minnesota, North Dakota and South Dakota
KC	New England, including Maine, Massachusetts, Connecticut, New Hampshire, Rhode Island and Vermont
KE	Mid-Atlantic, including New York and New Jersey
KI	Southeast, including Alabama, Georgia, Florida, Kentucky, North Carolina, South Carolina, Tennessee and Virginia
KG	Mid-Atlantic, including District of Columbia, Delaware, Maryland and Pennsylvania
KK	Gulf Coast, including Arkansas, Louisiana, Mississippi, New Mexico, Oklahoma and Texas
KM	West Coast, including California
KO	West, including, Arizona, Idaho, Montana, Nevada, Utah, Washington, Wyoming and Oregon
KQ	East central, including Ohio, Michigan and West Virginia
KS	Central, including Illinois, Indiana and Wisconsin
KU	Pacific, including Hawaii
KW	Alaska
WW	Puerto Rico

(if 35.22 is not used in your area, use 35.58). If you are using a tunable receiver the calibration will probably not be accurate enough. In this case, you'll have to mark your dial for these channels when skip, and pagers, are in.

Once you've set your dial on 35.22 or 35.58 MHz, turn the squelch up slightly and leave it on. When conditions start to pick up, you'll hear activity. Now flip up to the 43-MHz channels and lock in. If activity is not noted on these channels within the next hour or so, periodically check back to the 35-MHz channels. The muf may never have made it much beyond 35 MHz.

On the other hand, the Es may be into an area with no 43-MHz pagers (or worse yet, with no pagers!). So, when 35 MHz is active, alert yourself to a *possible* sporadic-E opening on six meters. When 43 MHz is active, watch for a *probable* six meter Es opening within the next few minutes!

If you're using a scanning receiver, and one that will cover an eight MHz stretch on low band, let it scan all the paging channels open in your area. If it sticks on one of the 35-MHz channels, switch that position off, and let the receiver scan the 43 MHz channels only. With some practice you'll start to get an idea when six-meter Es is likely.

General directions of the Es origin can be plotted for the expected six-meter opening. As you might expect, the 43-MHz channels are of greatest help for this task; 35 MHz is of less use as refractive angles are substantially sharper. In addition, ionization patches that affect 35 MHz may allow 50 MHz or 43 MHz to pass.

Actually, there are an almost infinite number of stations operating within this frequency range. When Es is strong, numerous public service stations and other two-way stations pour in, adding useful information about the potential opening. A mobile phone channel around 35.35 MHz is also very helpful, as mobile radiotelephone operators identify by their cities. Europeon tv audio channels can be monitored as well. BBC audio is transmitted on 41.5 MHz (BBC-1 service is scheduled to be discontinued before the next sunspot maximum) and the French ORTF is on 41.25 MHz.

During the recent sunspot peak, BBC and ORTF were both received on many occasions across the US. BBC-1 has also been received on the East Coast via auroral E. Over most of that sunspot peak, the 30- to 50-MHz range was quite lively. Paging stations on 35-MHz were received here in Chicago on a daily basis from Hawaii, California, Puerto Rico and other locations. Of course, changes in sunspot activity affect reception greatly, often creating a mass of hetrodynes on the paging channels. Most common here were Fresno, California and San Juan, Puerto Rico, both on 35.22 MHz.

Armed with this tool for catching six-meter Es openings you're set for more vhf fun than you ever bargained for!

references

1. Bob Cooper, KV4FU, "An Automatic Two-Way DX Beacon for VHF," *ham radio*, October, 1969, page 52.
2. Ed. Noll, W3FQJ, "VHF Beacons," *ham radio*, December, 1971, page 66.

ham radio

low swr dipole pairs for 1.8 through 30 MHz

A simple antenna system for the hf bands that offers low swr and broad bandwidth without a tuner

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After reading **Bill Orr's article** on multiband dipoles for 10, 15 and 20 meters, I decided to try some of his ideas on the lower-frequency bands, together with a few innovations needed to solve the bandwidth problem.[1] If you calculate the percentage bandwidth of each amateur band from 160 through 10 meters, you come up with the figures in **table 1**.

Reference to the table tells us immediately that the only amateur bands requiring broadband treatment are 80 and possibly 10 meters, since conventional dipoles will readily meet the bandwidth and swr requirements for bandwidths of 5% or less.

The first dipole pair to be considered in an all-band installation is the 160- and 80-meter combination, since it is the biggest and is usually mounted near the top of the mast or tower. Both antennas may be mounted as inverted Vs as long as a minimum angle of thirty degrees is maintained between them. The thirty degree separation is necessary to keep antenna interaction to a minimum.

The 160-meter section is a simple wire dipole cut for 1850 kHz and then adjusted for minimum swr with the 80-meter dipole connected and mounted in place.

The 80-meter dipole is a standard broadband double bazooka.[2,3] This antenna has a fairly good broadband characteristic and works well with the 160-meter dipole as a two-band antenna system.

Construction details for my antenna are shown in **fig. 1** and **tables 2** and **3**. They depart somewhat from the original design, but mine are simpler and perform

table 1. Bandwidths of the hf amateur bands. Percentages were determined by dividing the bandwidth in MHz by the center frequency.

band	bandwidth (percent)
160	5.1
80	13.3
40	4.2
20	2.7
15	2.1
10	5.9

just as well. I used twinlead for the ends instead of ladder line because twinlead was more readily available.

The antenna is most easily set up by cutting the coaxial center section for 3750 kHz according to the formula in **table 3** and then adjusting the length of the end sections for *identical swr at each* band edge. This final adjustment must be made with the antenna connected to the 160-meter dipole. The result should be an swr curve similar to **fig. 2** for both dipoles. If the 2:1 swr on the band edges is objectionable, the curve may be modified by lengthening or shortening the twinlead ends slightly to favor that portion of the band used most frequently.

The double bazooka is a fine antenna for this application — but, as usual, there is a catch. It works well in conjunction with any dipole that is *lower* in frequency than the double bazooka. Thus, a 10-meter double bazooka will work well with a 20-meter dipole, just as an 80-meter double bazooka will work with a 160-meter dipole. However, any attempt to use the antenna with a *higher* frequency dipole will result in an absolutely unmanageable swr on the higher-frequency band. This peculiar behavior is caused by the center conductor of the *coaxial section acting as a pair of* short-circuited quarter-wave stubs on the second harmonic. Therefore, if 160-meter operation is not desired, the 80-meter double bazooka will have to be run on a transmission line separate from the other dipole pairs.

40 and 15 meters

Well accepted theory states that dipoles working from a common feedline or balun must be harmonically related to be effective, the implication being that the relationship usually involves the fundamental and the second harmonic. The question arose, when developing these dipole pairs, what to do about fifteen meters? The only amateur band to which

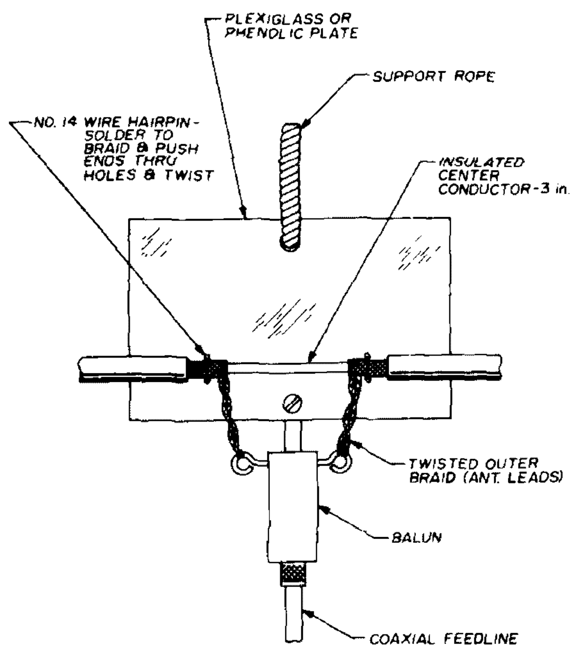
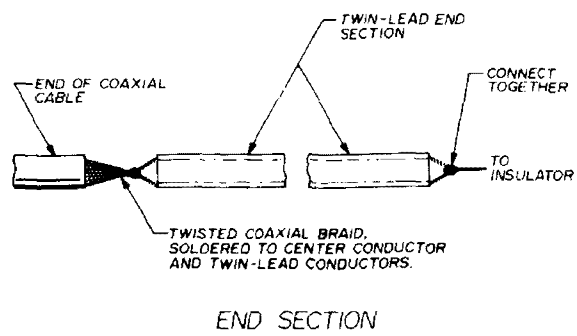


fig. 1. Construction details for the 80- and 10-meter broadband dipoles. Table 3 includes the necessary formulas and some typical dimensions. The end segments (D3) will vary somewhat from the calculated values and, like all dipoles, may need *pruning once the* antenna is in place.



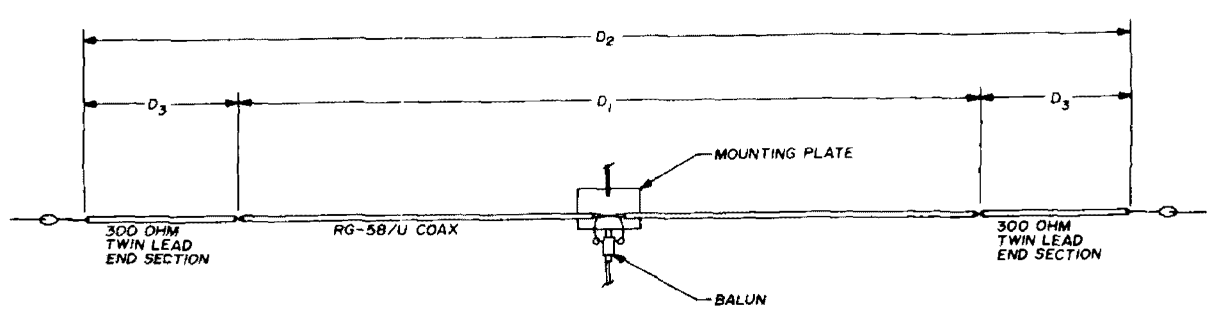


table 2. Wire dipole lengths.

band	overall length
160	243' 0"
40	64' 2"
20	33' 0"
15	22' 0"

it is harmonically related is forty, so I set up a 40-meter dipole with a low swr. When I tried operating on fifteen, the swr jumped to 3:1 and higher. This should really not be surprising since theory again tells us that the radiation resistance of an antenna increases as the number of half waves increases. The thought then occurred that the addition of a 15-meter dipole to the 40-meter dipole might solve the problem. This worked out very well indeed, as a reference to fig. 3 will reveal.

The 40- and 15-meter combination, unlike the previous dipole pair, does not require the application of broadband techniques. Conventional single-wire dipoles will do the job and, if carefully adjusted, will result in a swr of 1.5:1 or less on both bands. Incidentally, I did not attempt to center the swr curves as in the previous case. The 40-meter dipole is a little short and the 15-meter dipole a little too long, a situation readily corrected by the perfectionist.

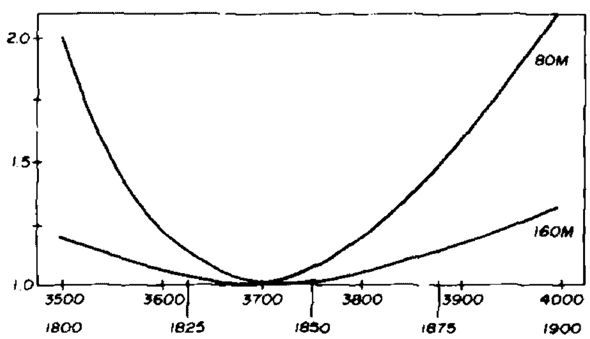


fig. 2. Swr curves for the 160- and 80- meter dipole pair. Dimensions of the final antenna lengths are given in tables 2 and 3.

To set up the pair, begin by adjusting the 40-meter dipole for equal swr at the band edges. This should be done with a pre-cut 15-meter dipole connected to the 40-meter dipole. Also, be sure to space the ends of the 15-meter dipole at least 8 feet from the 40-meter dipole, otherwise severe antenna interaction may result. This will show up as an excessive swr, usually on 15 meters. After obtaining a satisfactory swr curve on forty, proceed to adjust the 15-meter dipole. Usually very little adjustment of the 15-meter element will be required.

20 and 10 meters

The 10- and 20-meter dipole combination may be regarded as a scaled-down version of the 160- and 80-meter dipole pair, since both pairs use a simple wire dipole for the lower-frequency band and a broadband antenna for the higher-frequency band. The similarity ends there. You will find that the dipoles of this antenna pair are more interdependent and therefore more critical to adjust.

Begin by cutting both antennas to their calculated length. Connect both to a common balun, and make sure that the ends of the 10-meter dipole are at least 10 feet from the 20-meter dipole.

Adjust the 20-meter antenna for minimum swr. After obtaining a satisfactory swr similar to fig. 4, proceed to adjust the 10-meter dipole for minimum swr by adjusting the lengths of the twinlead ends. As a final touch, check both bands once again to see if any significant shift has occurred in either swr curve. By following this procedure you will experience little difficulty in setting up this last dipole pair.

The dipole pairs described in this article are the direct result of much experimentation done while developing a

table 3. Bazooka antenna dimensions. D1, D2 and D3 are shown in fig. 1. The formulas used to compute these lengths initially are D1=492/f(MHz) for foam coax and D1=325/f(MHz) for polyethylene coax. D2=460/f(MHz) and D3=(D2D1)/2.

band	D1	D2	D3	cable
80	105'	125'	10'	foam
10	11' 2"	15' 6"	26"	poly

simple, effective, all-band antenna system for Doc, WB6UWK. Doc is a blind amateur who likes to work all bands on all modes without the need for fooling

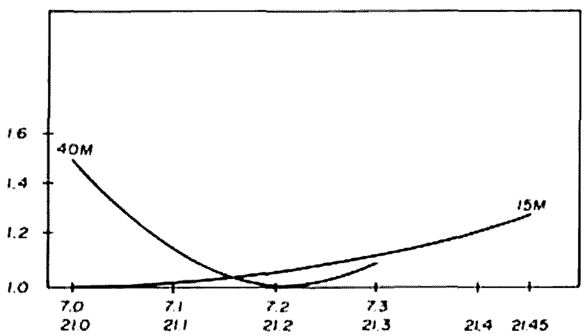


fig. 3. Swr curve for the 40- and 15- meter dipole pair.

with couplers or matching gimmicks of any kind. A coaxial switch makes antenna switching extremely simple.

The dipole pairs have been in service for about a year now and have given no trouble, even when running the legal

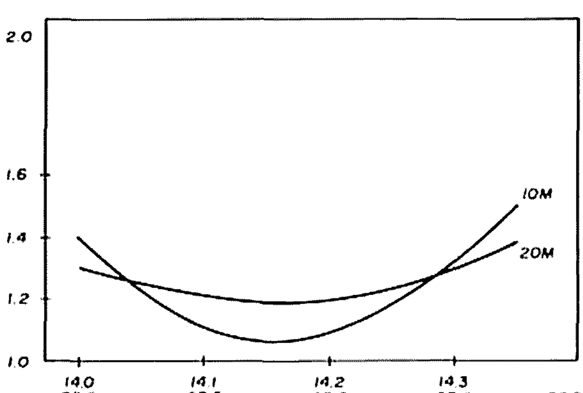


fig. 4. Swr curve for the 20- and 10- meter dipole pair.

power limit. No superlative DX claims are made for this system, but on each band it seems to perform as well as a simple dipole.

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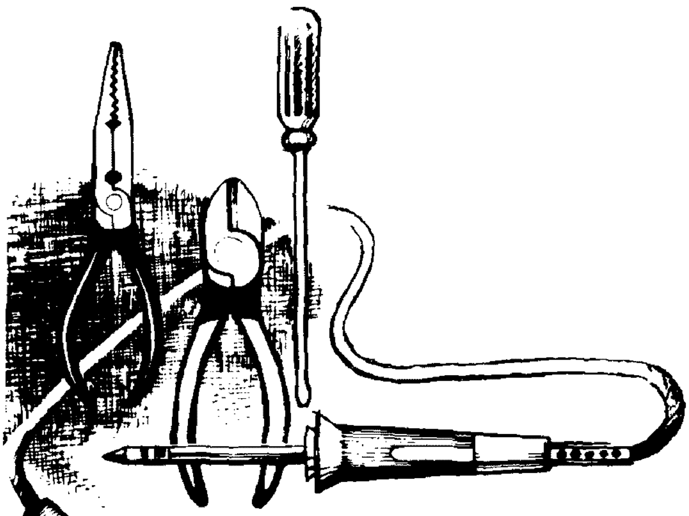
1. William I. Orr, W6SAI, "Multiband Dipoles for Portable Use," *ham radio*, May, 1970, page 12.
2. "The Radio Amateur's Handbook," 48th edition, ARRL, 1971, pages 368-369.
3. Charles C. Whysall, W8TV, "The Double Bazooka Antenna," *QST*, July, 1968, page 38.

ham radio



circuits and techniques

ed noll, W3FQJ

multi-function ICs

Multi-function integrated circuits comprise two or more basic functional systems within the same case. Input, output and control leads from each system are brought out. Hence, the internal systems can be interwired externally to serve a variety of applications. These versatile devices are a specialty of the Exar Corporation. Last month the XR-205 Waveform Generator was introduced. Several additional circuits of this 16-pin in-line device are described this month. Also covered is their multi-function XR-S200, mounted in a 24-pin, in-line package. This is an extraordinary unit planned for communications systems.

XR-205 applications

In the XR-205 circuits covered previously the upper frequency limit was no more than several megahertz. A modified circuit, **fig. 1**, permits a sinusoidal output up to at least 10 MHz. Output level at pin 11 is less in this circuit arrangement being approximately 700 mV. Frequency capability is now in the amateur vfo spectrum for either high-frequency or vhf operation. This single device can serve as a vfo (keyed or unkeyed), double-sideband a-m, suppressed carrier (for single-sideband application), or fm signal source. All are tied together in one neat package. Furthermore, it makes available a variety of non-sinusoid signals for transmitter test circuits, synchronizing or other applications.

Such a basic signal source could be

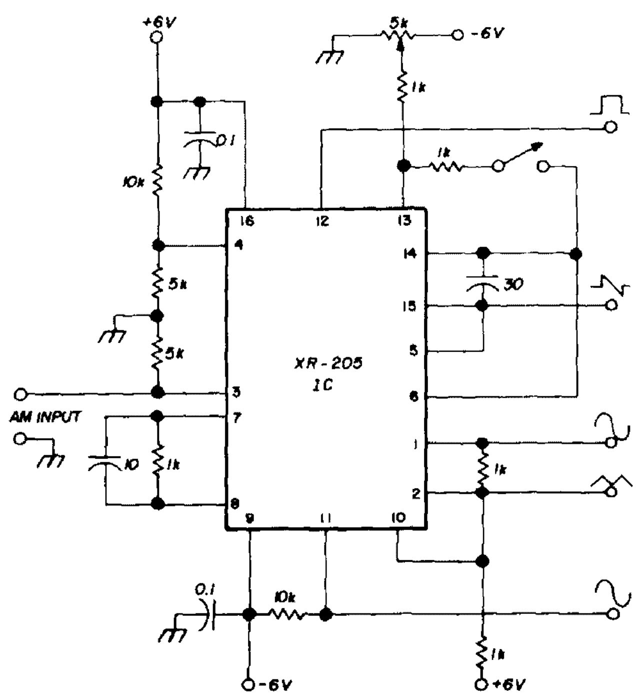


fig. 1. High-frequency operation of the Exar XR-205 waveform generator IC.

followed by a chain of linear amplifier and mixer-oscillator integrated circuit stages, fig. 2. The final power amplifiers of such a transmitter could be designed to operate either in the linear or class-C

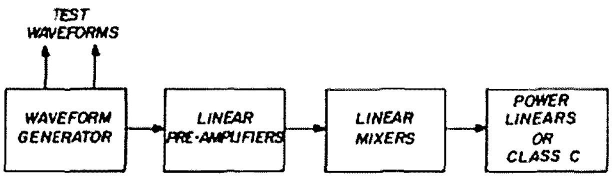


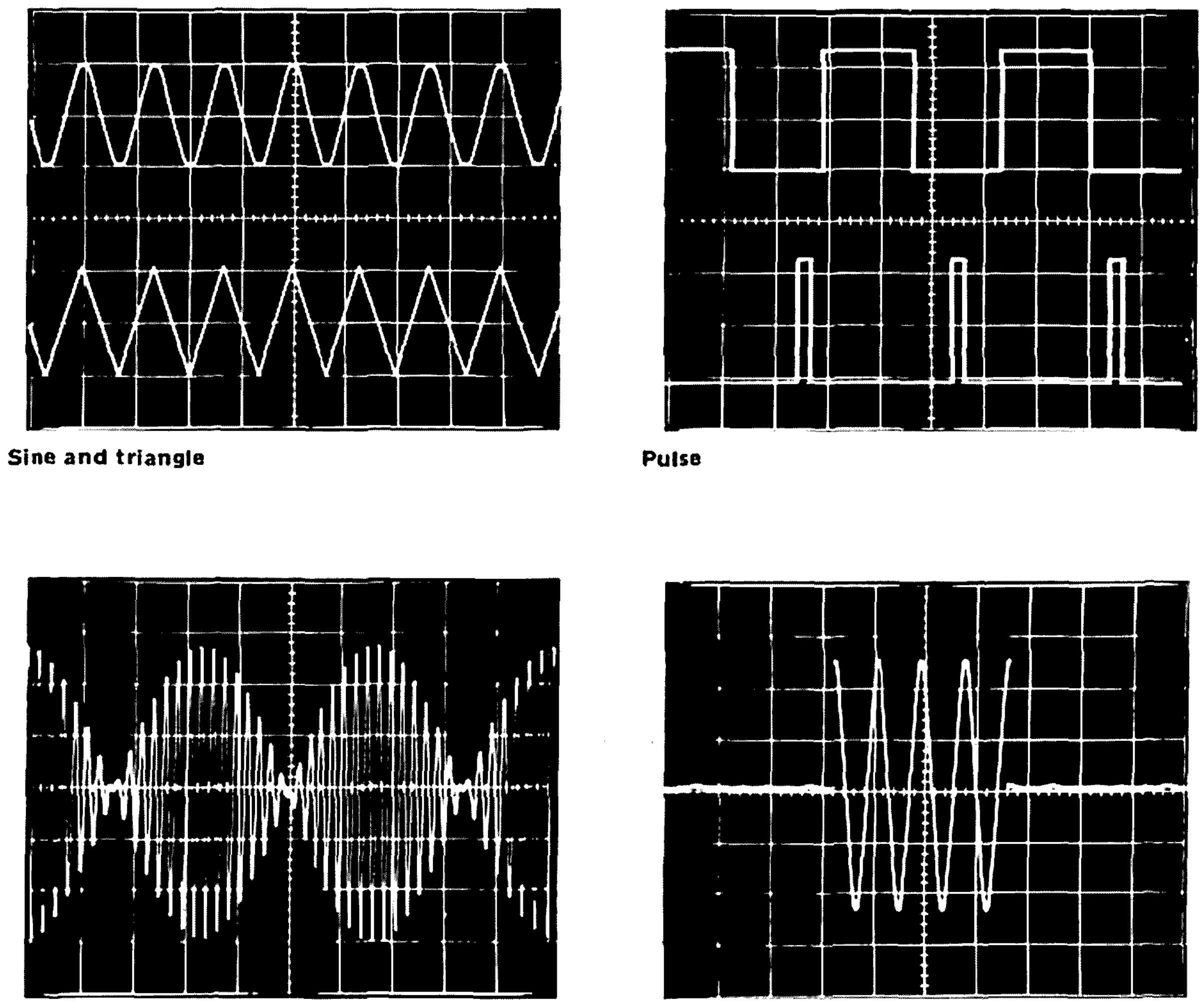
fig. 2. Block diagram of a transmitter using the waveform generator IC.

amplifier mode. The latter would permit more efficient CW and fm signal amplification when desired. Several basic and modulated wave-forms are given in fig. 3. Also, the circuit for single-supply operation is given in fig. 4. This circuit has an upper frequency limit of 2 to 4 MHz and an output voltage of 2 to 3 volts peak-to-peak. Characteristics are given in table 1.

The single unit cost of $16.00 might jar you. But how much would it cost to duplicate this versatility with discrete transistors? What a tremendous fortune it would take to duplicate this performance with vacuum tubes!

Exar makes available an a-m/fm generator design kit for $28.00. This includes two XR-205 generators, a printed circuit board (etched and drilled), detailed components parts list, and assembly instructions. Estimated cost of additional parts is approximately $27.00.

The basic circuit is given in fig. 5. The IC on the left provides a modulating signal. Sine, square, triangle or ramp type is selected with switch S1. Frequency of



Sine and triangle

Pulse

Suppressed-carrier a-m

CW

fig. 3. Typical waveforms generated by the XR-205 IC.

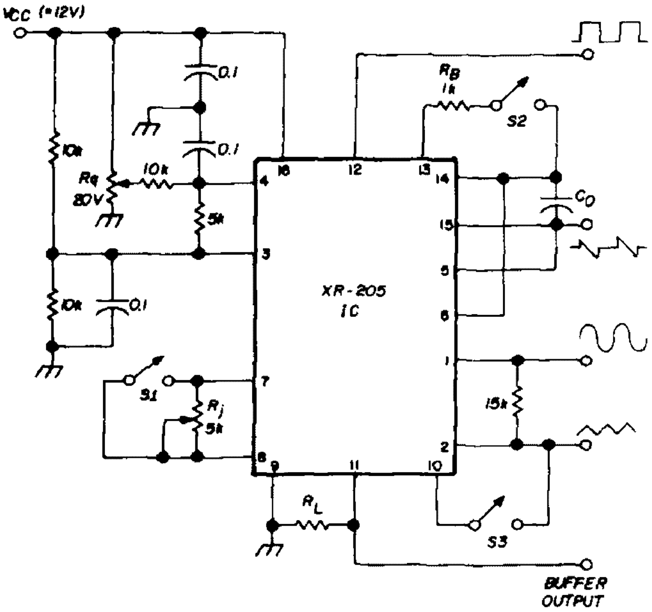


fig. 4. Test circuit for single-supply operation of the XR-205.

operation is set by capacitor C1. A combination of switched and variable capacitors at this point permits adjustment of modulating wave frequency. In the formation of non-sinusoidal modulating waves switch S2 selects a duty cycle of either 20 or 50 percent. Switch S3 is the modulation mode selector, fm, a-m or CW.

The second waveform generator operates at high frequency. It includes switches S4 for setting the duty cycle of non-sinusoidal output, and S5, which determines the output waveform. Capacitor C2 determines the output frequency. Again, switched capacitors and a variable capacitor permit frequency control. The adjustment of the modulation and carrier waveforms can be made with potentiometers R4 and R5. Potentiometer R1 sets the level of the modulating signal while potentiometer R3 sets the output carrier level. Potentiometer R2 is a carrier zero adjustment for use with suppressed-carrier modulation.

multi-function XR-S200 IC

The XR-S200 is also a three-function unit which includes an analog multiplier, vco and operational amplifier. More versatility, including operation as phase-locked loop (PLL), is present as compared to the XR-205, which consists of a voltage-controlled oscillator, balanced modulator and buffer amplifier. A number of additional applications are possible, and fre-

table 1. Characteristics of the XR205 circuit shown in fig. 4.

	min	typ	max		
Sinusoidal:					
Upper Frequency Limit	2	4		MHz	Measured at Pin 11
Peak Output Swing	2	3		V pp	S1, S3 closed
Distortion (THD)		2.5	4	%	S2 open
Triangle:					
Peak Swing	2	3		V pp	S1, S2 open
Non-Linearity		±1		%	S3 closed
Asymmetry		±1		%	f = 10 kHz
Sawtooth:					
Peak Swing	2	3		V pp	S2 open
Non-Linearity		1.5		%	S2 and S3 closed
Ramp:					
Peak Swing	1	1.4		V pp	S2 and S3 open pin 10
Non-Linearity		1		%	shorted to pin 15
Squarewave (Low Level):					
Output Swing	0.5	0.7		V pp	S2 and S3 open, pin
Duty Cycle Asymmetry		±1	±4	%	10 shorted to pin 12
Rise Time		20		ns	10 pF connected from
Fall Time		20		ns	pin 11 to ground

quency range is extended to 30 or 40 MHz. The additional versatility requires a 24-pin case and unit cost of $28.00.

Applications for the XR-S200 include phase-locked loops, fm demodulation, FSK detection, PSK demodulation, signal conditioning, tracking filters, frequency synthesis, telemetry coding and decoding, a-m detection, (quadrature and synchronous detectors), linear sweep and fm generation, tone generation and detection, waveform generation and analog multiplication. That's quite a list, and you say there is no more room for experimentation in ham radio?

How many applications for amateur radio use could you come up with using this device? Maybe it would help us to be frank and admit that often we don't want to assign the time, or we have lost or never developed the patience and perseverance that experimentation requires.

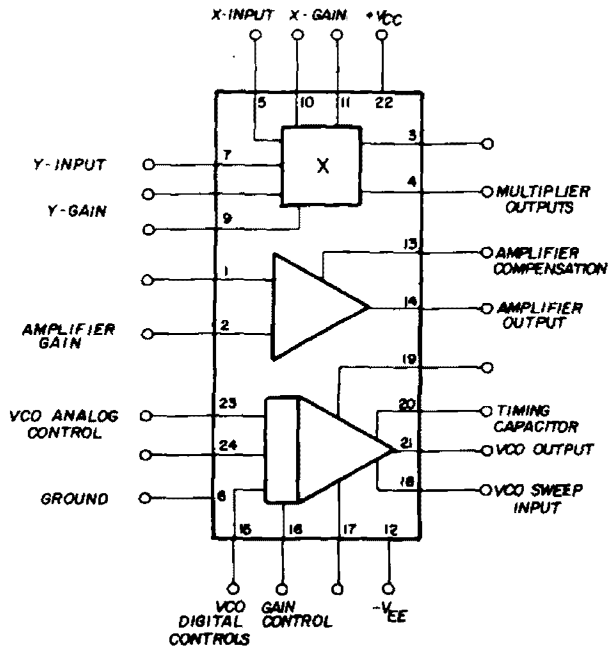


fig. 6. Diagram of the multi-function Exar XR-S200 IC.

typical circuits

The pin-out diagram for the multiplier, amplifier and vco is given in fig. 6. The analog multiplier is an especially versatile section which can be used as detector, balanced modulator/demodulator, frequency multiplier, etc. Several external components permit you to assemble a phase-comparator very simply as shown in fig. 7. You will obtain an output voltage that corresponds to the relative phase of reference and signal inputs.

The three quite similar schematics of fig. 8 show how the multiplier section can be used as suppressed-carrier modulator,

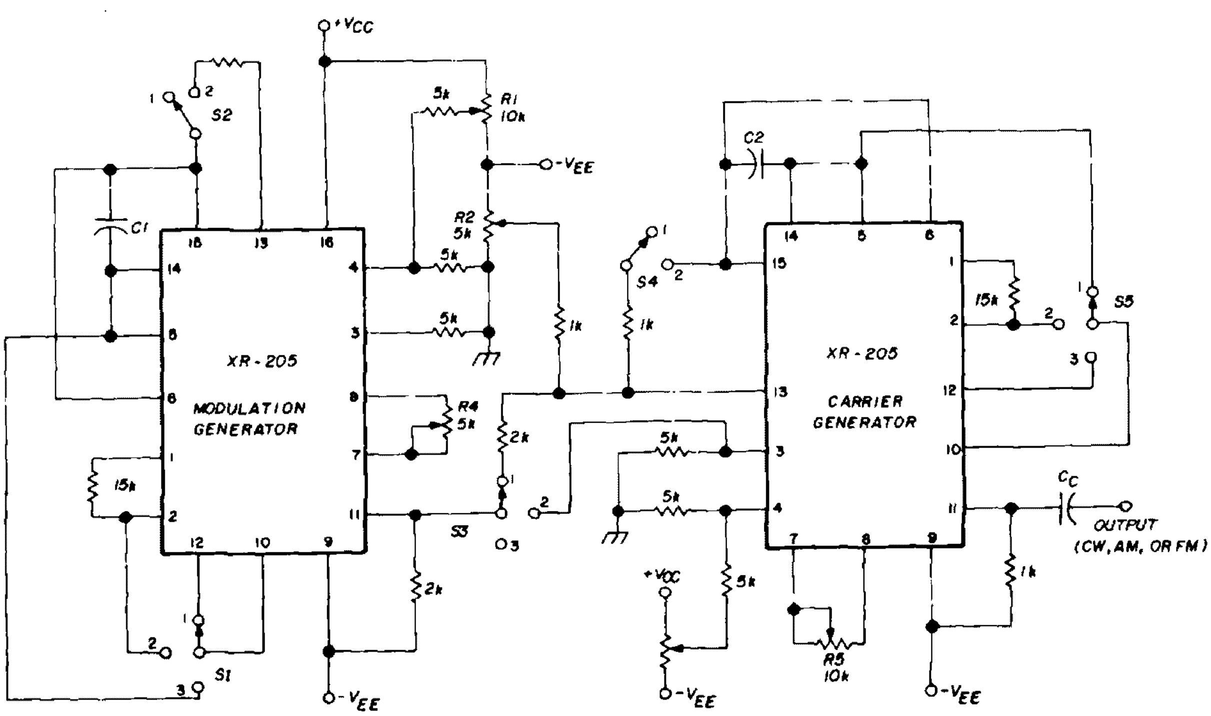


fig. 5. Self-contained a-m/fm generator using two XR-205 ICs.

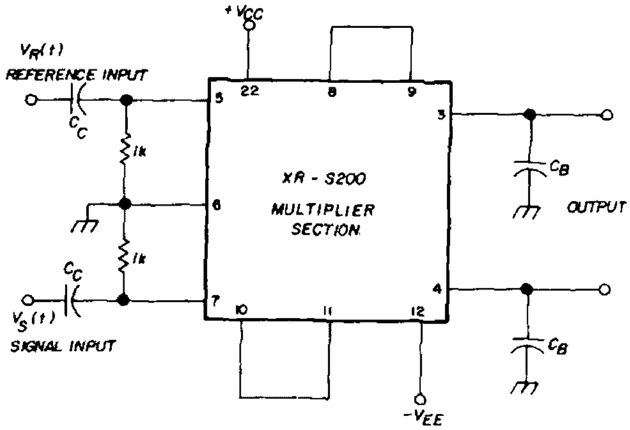


fig. 7. Phase comparator circuit based on the multiplier section of the XR-S200 multifunction IC.

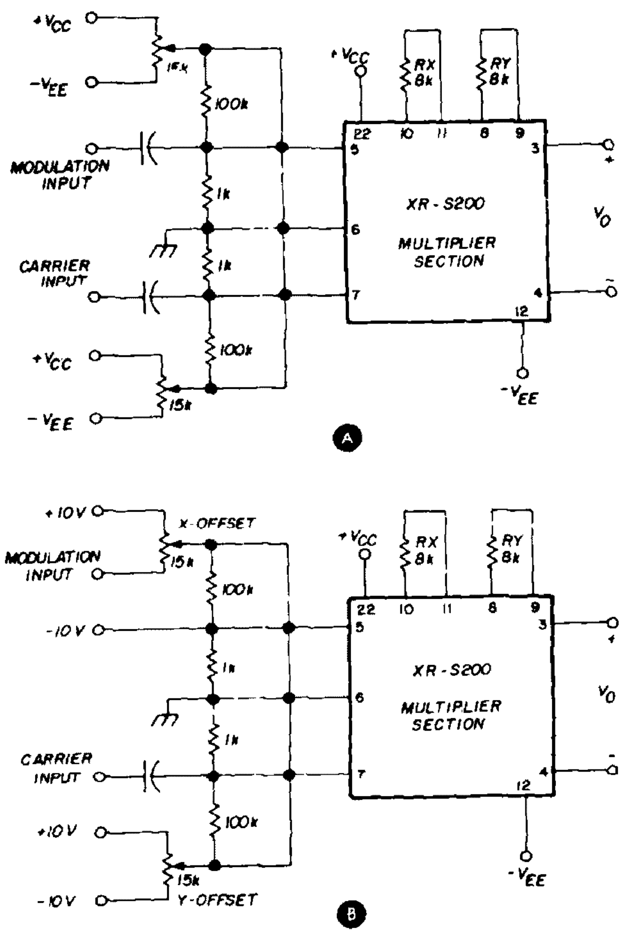
double-sideband a-m modulator and frequency doubler. Your source of carrier signal can be the vco section. Come up with a simple switching arrangement and you will have a versatile CW, a-m and dsb signal source. Inasmuch as the vco can be frequency modulated, you will also have an fm signal source. Since the vco can also be crystal controlled you can generate a signal source with crystal stability. The simple circuit of **fig. 9** will do the job.

The device also has receiver applications and, therefore, versatile transceiver possibilities. A simple a-m detector using only the multiplier is given in **fig. 10**. The PLL connection arrangement, **fig. 11**, is appropriate for demodulating CW, ssb, dsb and FSK. A PLL fm demodulation system, **fig. 12**, uses the multiplier section as the phase detector. The voltage-controlled oscillator and external resistor-capacitor filter provides fm demodulation. The amplifier section increases the level of the demodulated audio.

There is more. The same device can be used as a waveform generator, forming the same variety of signals possible with the XR-205.

three-sided antennas

A letter from Jim Gray, W2EUQ, indicates that he has become an eager three-sided antenna experimenter. His open-loop configuration is shown in **fig. 13.** He corner-feeds at the apex. This differs from the usual triangle and its feed point at the center of one of the sides. The antenna operates 5/2-wavelength on 15 meters, and 3/2-wavelength on 20. Jim uses a knife switch to add an extension for proper resonance on 14 MHz.



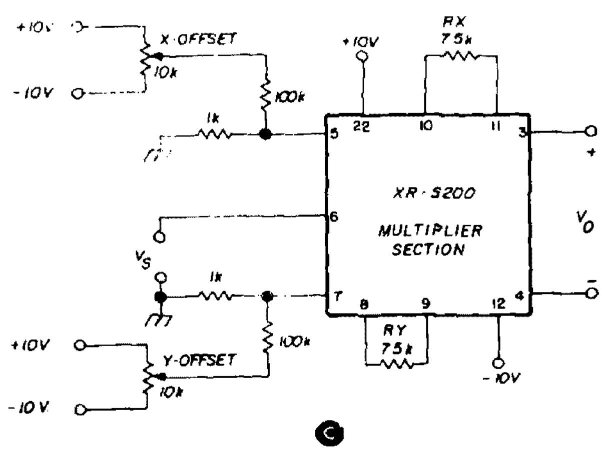


fig. 8. Using the multiplier section of the XR-S200 as a suppressed-carrier a-m modulator (A), double-sideband a-m modulator (B) and frequency doubler (C).

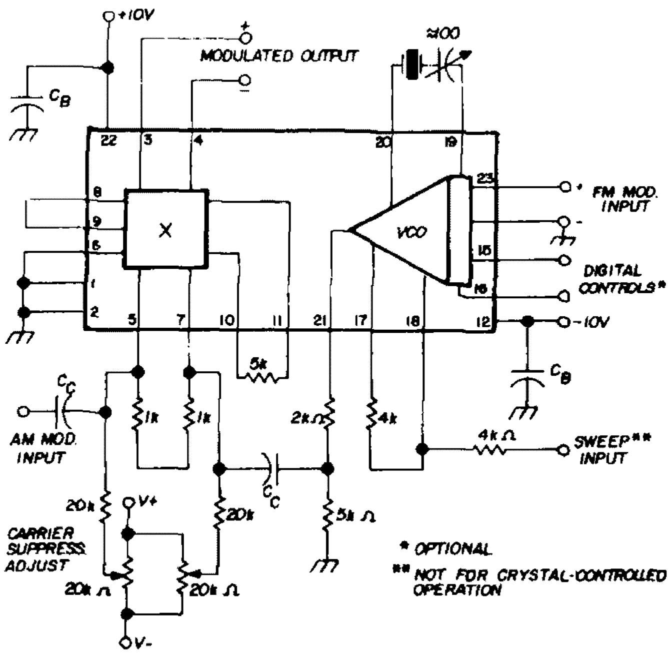


fig. 9. Circuit diagram of an a-m and fm signal source based on the XR-5200 integrated circuit.

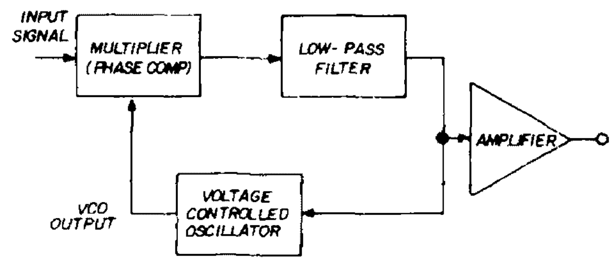


fig. 11. Phased lock loop is suitable for demodulating CW, a-m, ssb, dsb and FSK.

When the end of the extension is shorted, the assembly becomes a closed full-wave loop on 40 meters. In this case it operates as a delta loop because of the corner feed point.

Another three-sided arrangement comes from the South Australian Wireless Institute Journal by way of Leo Gunther, VK7RG. This 160-meter antenna was designed by VK5EF for a site with space limitations. As shown in fig. 14 it is fundamentally an 80-meter inverted dipole with two sides folded around toward the mast. These include an 80-meter trap and an appropriate wire extension to obtain 160-meter resonance. VK5EF resonates his trap on 3.6 MHz. With the dimensions given there is also short-length resonance on 1.82 MHz. No antenna tuner is required.

interference-reducing antennas

The catv industry has been doing much research in the development of co-channel interference-reducing antenna systems. Two techniques that have possible application in radio amateur circles appeared in the June, 1972 issue of *Communications News.*[3,4] The work is

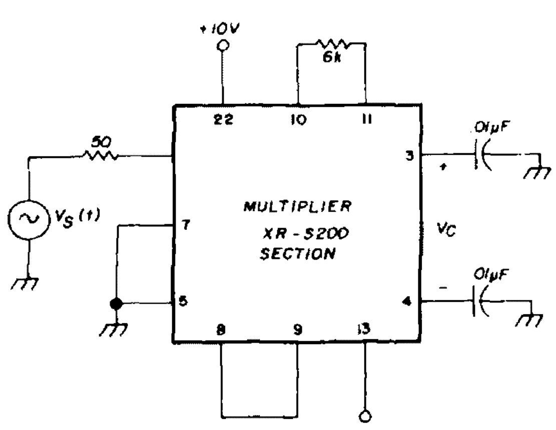


fig. 10. A-m detector circuit.

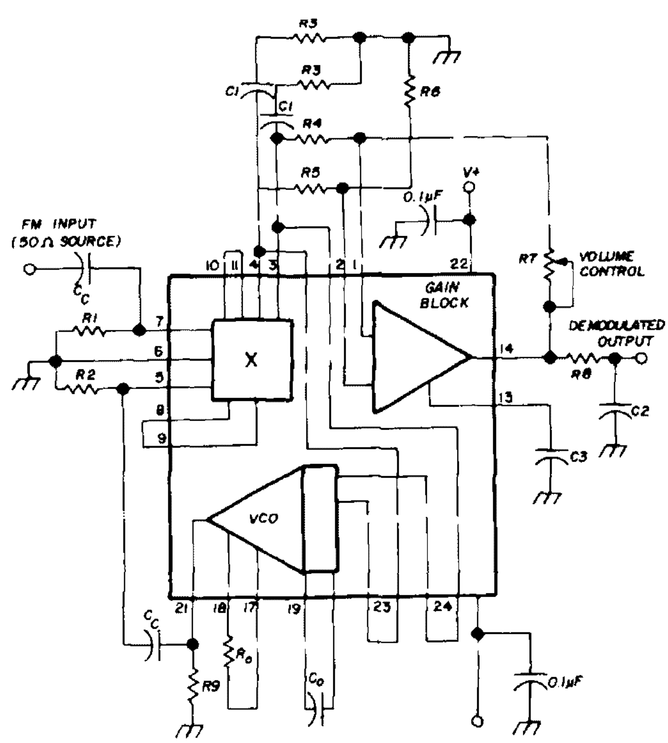


fig. 12. Fm demodulator using the phased-lock loop.

being conducted at Scala Radio by its president Bruno Zucconi and project engineer Charles B. Carter. An improvement in front-to-back ratio is accomplished by placing the top antenna one-quarter wavelength in front of the bottom one as shown in fig. 15. This top

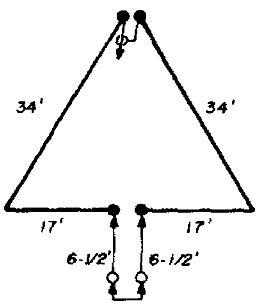


fig. 13. Three-sided antenna designed by W2EUQ for use on both 15 and 20 meters. The 6½-foot extension is switched in with a knife switch to provide resonance on 20 meters.

antenna is also fed 90° later than the bottom one. Table 2 provides useful dimensions for the amateur bands from two through twenty meters. Values are for band center.

If your interference is largely from the forward direction, two Yagis or other narrow-beam, high-gain antennas are fed in phase, fig. 16. With suitable horizontal displacement there is signal cancellation from some specific receive angle off the beam direction. You only have to know the angle of arrival of your most troublesome QRM. The information in table 3 then permits you to so space the two

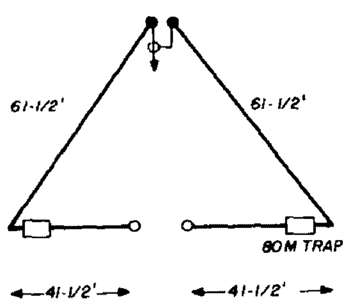


fig. 14. This 160-meter loaded delta antenna built by VK5EF is used on both 80 and 160 meters.

table 2. Antenna spacing and feedline length for increasing front-to-back ratio of stacked Yagi antennas.

band	2 meters	6 meters	10 meters	15 meters	20 meters
¼-wave free space	20¼"	57"	8'6"	11'7"	17'4"
¼-wave line*	13¼"	37½"	5'8"	7'7"	11'6"

*based on 0.66 velocity factor.

antennas to give the most rejection at a particular angle off the forward beam direction.

The pattern of fig. 17 shows the type of deep null that can be obtained 20° off the beam direction. With your antenna system mounted on a rotator, you can do some fine nulling of interference coming from the same general direction. Perhaps

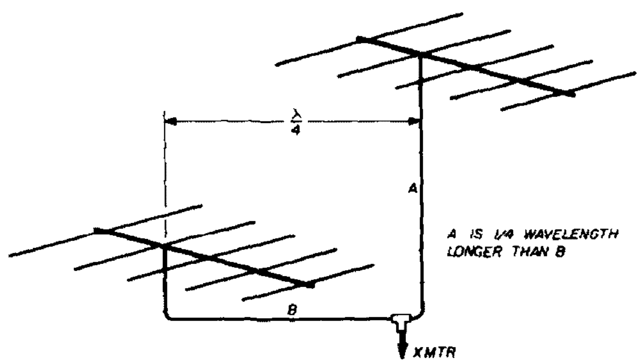


fig. 15. Improving the front-to-back ratio of stacked Yagi antennas with offset vertical mounting.

20 degrees is not the ideal angle in amateur operation. Some experimentation with horizontal displacement of the two Yagis may help you decide on the most favorable null angle for your particular location.

By the laws of reciprocity your trans-

table 3. Spacing of Yagi antennas to obtain desired null angles as described in the text.

null angle	10°	15°	20°	25°	30°	35°	40°	45°	50°	55°	60°
spacing in wavelengths	2.5	1.75	1.5	1.25	1.0	.85	.75	2.25	1.9	1.7	1.65

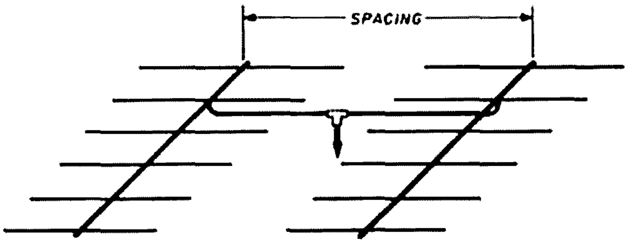


fig. 16. Desired null angle as shown in Table 3 is determined by the spacing between the two Yagi antennas.

mit pattern will be similar. On 2-meter fm these techniques have great possibility in areas crowded with repeaters operating on the same and adjacent channels. Your antenna could be made to beam on your particular repeater, and still have nulls in the directions of the repeaters you do not wish to activate. Operate your base sta-

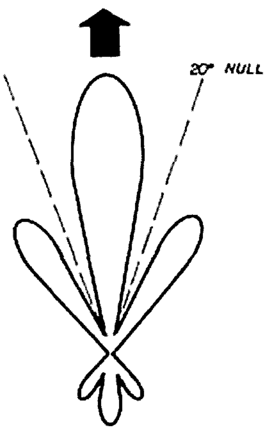


fig. 17. Response of two five-element Yagi antennas positioned for a 20° pattern null.

tion with two antennas, one omnidirectional for simplex and one highly directional for repeater communications.

references

1. "XR-205 Monolithic Waveform Generator Kit," Exar, 733 North Pastoriz Avenue, Sunnyvale, California 94086.
2. "XR-S200 Multi-Function Integrated Circuit Specifications and Application Notes," Exar, 733 North Pastoriz Avenue, Sunnyvale, California 94086.
3. Charles B. Carter, "Increasing Front-to-Back Ratio," *Communications News,* June, 1972, page 56.
4. Bruno Zucconi, "Reducing Co-Channel Interference by Spacing Antennas on Same Plane," *Communications News,* June, 1972, page 54.

ham radio

printed-circuit board

for the RTTY electronic speed converter

A printed-circuit board is now available for this versatile and accurate RTTY speed converter originally described by WA6JYJ

Earl E. Palmer, W7POG

The RTTY electronic speed converter described by WA6JYJ in a recent issue of *ham radio*[1] provides an inexpensive method of converting low-speed RTTY signals for printout on a higher speed teleprinter. This is much more simple than the mechanical gear shifts that are usually required for this purpose.

The circuit, which uses ten ICs, seven transistors and a bunch of other components, is quite sophisticated and strains the ability of the most proficient home builder. I attempted to simplify construction by replacing the point-to-point wiring with a printed-circuit board. With the exception of rearranging the IC pin numbers to simplify the board layout, this was easily accomplished. Drilled printed-circuit boards are now being made available to other interested RTTY amateurs.*

construction

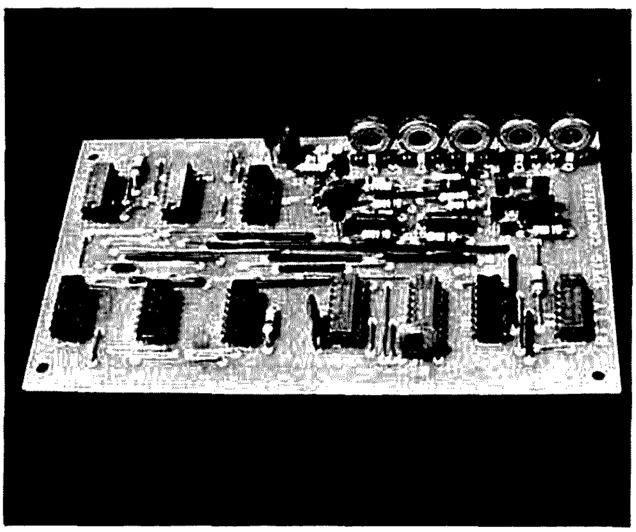
To keep construction costs down, a single-sided printed-circuit board is used. This requires jumpers between some of the pads on the board. These jumpers should be made with insulated, tinned number-20 hookup wire. Complete instructions are furnished with the boards.

The board is designed for inexpensive, vertically-mounted 50,000-ohm trimmers. These are used to set the unijunction *oscillator frequency.* To install the miniature 20-turn trimmers suggested by WA6JYJ, the board must be re-drilled to accept their terminals, and jumpers installed to make the necessary connections.

*Drilled, fiberglass printed-circuit boards for the RTTY speed converter *are available from* P&M Electronics, Inc., 519 South Austin, Seattle, Washington 98108. The price is $6.00. The board *includes* a schematic of the converter as well as a component layout diagram.

The unijunction oscillator frequency is determined by R1, C2 and the 50k trimmer (see fig. 1 of WA6JYJ's article). With the inexpensive board-mounted trimmers I used (CTS type Z-201-R503B), a 10% Mylar capacitor and 5% deposited-carbon resistor resulted in satisfactory long-term frequency stability for the normal ambient temperatures found in most ham shacks. If the speed converter must be installed where wider temperature variations are encountered, precision components should be used.

The following suggestions are the results of building two separate speed converters. First of all, buy 100% tested integrated circuits. They cost a little more, between 10 and 50 cents each, but it's money well spent. ICs that are 100% tested are marked by the manufacturer with a "T" suffix or some other means, and must be specified when they are purchased.



Printed-circuit RTTY speed converter built by W7POG. Printed-circuit boards are available to interested readers.

Many manufacturers only sample test their ICs, and a few bad ones are bound to turn up in any production run. Of the ten ICs I purchased for my original speed converter, one was bad (a shift register), and many unnecessary hours were spent in locating the problem.

Also, when building the speed converter, be generous with noise-suppression capacitors. Noise problems exhibit themselves as unexplained, consistent garbles which defy troubleshooting. In addition, short leads to a well-regulated power supply are a must. The power supply suggested by WA6JYJ in his article is very satisfactory.

Two .01-μF noise-suppression capacitors are mounted on the printed-circuit board near each shift-register IC. If noise problems are still suspected, solder additional .01-μF, 600-volt disc ceramic bypass capacitors directly across the V_{cc} and V_{ee} IC pins on the copper side of the PC board. The shift registers and flip-flops are usually the source of noise problems.

When building the unit, be sure you use IC sockets. Use them if for no other reason than troubleshooting. The continuous-strip IC sockets manufactured by Molex are inexpensive and quite satisfactory for printed-circuit construction.

operation

When incorporating the RTTY speed converter into your station, make sure that the input of the converter is *after* the normal-reverse switch. The speed converter requires normal (not reverse) input for proper operation.

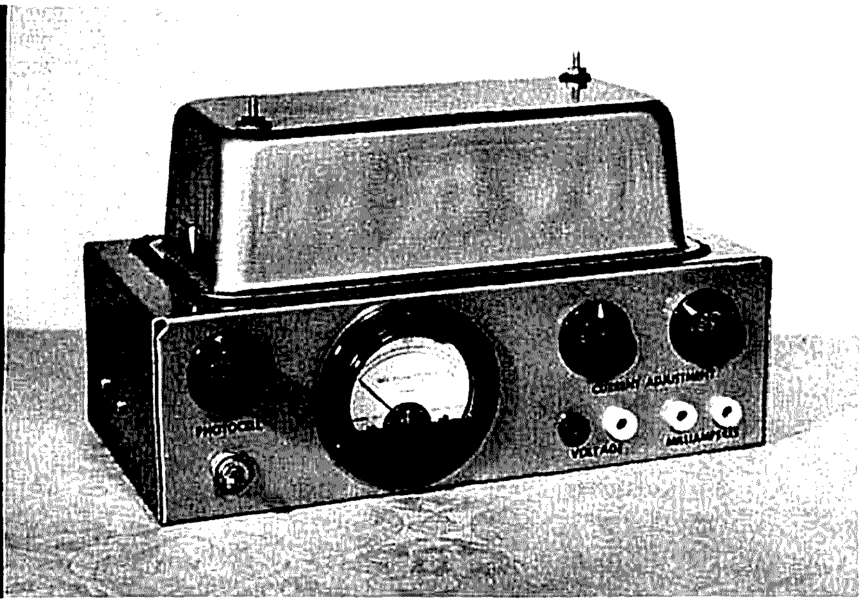
When snooping around for press copy, it's often difficult to determine what speed you are receiving. The best method I have found is to measure the period of the incoming bauds with an oscilloscope with a calibrated time base. For 67 wpm, the period is 20 milliseconds, slightly longer for 60 wpm, and slightly shorter for 75 wpm. The period for 100 wpm is 13 milliseconds.

Multiple-speed conversion greatly enhances the versatility of any RTTY station, and it is especially useful when looking around the amateur bands. More and more multiple-speed stations are now being heard as amateurs add the ability to operate at high speeds. As more RTTY enthusiasts learn of the advantages of higher speeds, amateur usage is expected to increase still further.

reference

1. L.H. Laitinen, WA6JYJ, "Electronic Speed Conversion for RTTY Teleprinters," *ham radio,* December, 1971, page 36.

ham radio



an accurate rf power meter for very low power experiments

This inexpensive meter will measure low power levels up to 148-MHz without expensive calibration equipment

Howard F. Burgess, W5WGF

If you are fascinated by the miles per watt statistics of very low-power communication or if you just enjoy building miniature transmitters, you will have a lot more fun if you really know how much power your little rig is putting out. Rf power is not easy to measure, and most of us have had our troubles even at moderate power levels. With power below a watt a good measurement is a real problem. There are many good designs for rf power meters, but most of them require a precision load resistor accurate at your operating frequency. In addition, a good rf voltmeter or another wattmeter is usually required to complete the calibration. Not many hams have access to these items.

My simple meter requires no special test equipment or calibration, and all of the parts can be found in any Radio Shack or similar store. I decided to drag an ancient technique out of the cobwebs and combine it with a few pieces of modern hardware. The results were even better than I had hoped for. The final model measures power as low as one milliwatt and works as well on two

meters as it does on eighty. With care the accuracy of the readings can approach 1%.

theory

For many years experimenters have known that a small lamp with a short filament will require the same amount of rf, dc or audio *power* to bring the filament to a given temperature. To make an accurate power measurement, load the transmitter into a suitable lamp and record the light level with a photocell and meter. Remove the rf and substitute enough metered dc to give the same light reading.

Yes, it is true that the bulb will probably never have a resistance even close to 52 ohms, and that the resistance will change as the power level changes. This is no great problem. Most transmitters have enough tuning range to match the bulb for maximum output. This same tuning that gives you maximum output will also tune out any small reactance which may be in the power meter. Just keep the lead from transmitter to power meter as short as possible.

circuit

Originally, I had visions of photo transistors, transistor dc amplifiers and expensive meters to get the sensitivity that I thought the unit should have. Happily the schematic in fig. 1 shows how simple life can be at times.

The cadmium-sulphide photocell acts as a light-sensitive resistor: the more light, the less resistance. The battery, two resistors and meter form a simple ohmmeter circuit to indicate the change of resistance in the photocell. With moderate light on the cell, the meter can be set to full scale (similar to zeroing an ohmmeter). With no light on the cell, the meter will read 0. The capacitor across the cell is just to bypass stray rf from the cell.

In the rf portion of the circuit, the power to be measured is connected to J1 and is fed to the load lamp through the 0.01 blocking capacitor. When dc is substituted for the rf, it is adjusted to the correct value with the 30- and 100-ohm variable resistors. The dc connections to the load lamp are through the rf chokes which keep the transmitter rf out of the battery and meter circuits.

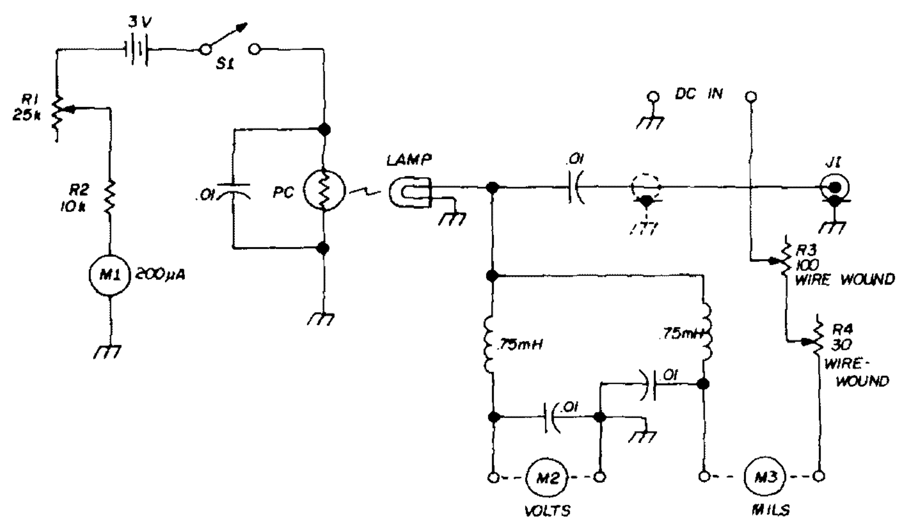
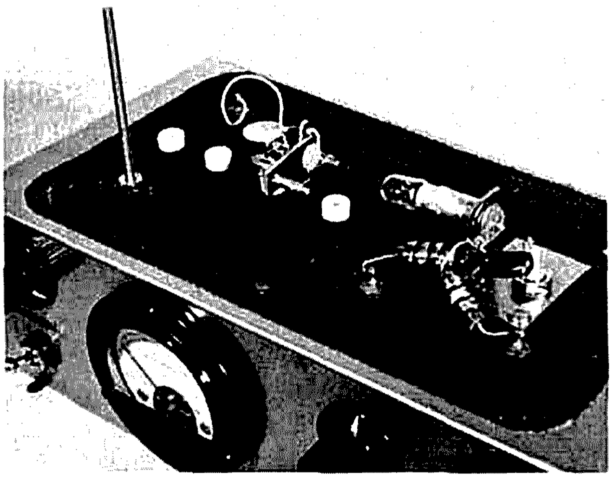


fig. 1. The schematic of the power meter.

The dc meters that are used to read the input power to the lamp were left external for convenience. In this way any meters which happen to be handy can be pressed into service. Meter M2 should be capable of reading the maximum voltage that will be used on the bulb, and M3 must be able to read the current required by the bulb at maximum voltage.

construction

I built the unit on an aluminum chassis, but it could just as well have been built on a piece of plywood. If you are not interested in the measurement of very low power, the system can be as simple as an old photo exposure meter and a load bulb. There are only two precautions to consider. Keep the rf leads short and heavy and enclose the photo cell and lamp in a light-tight compartment while making a measurement.



The photocell can be adjusted both for distance and angle by plugging the mount into the desired tip jack.

All of the parts are mounted on a 3 x 5 x 9-inch chassis. There is lots of empty space, but this makes life a little easier. A bread pan from the variety store makes a good chassis. The load lamp and photocell are mounted on top of the chassis to facilitate the bulb changing and photocell adjustment. The cover which forms a light tight compartment is a dime store cake pan. All surfaces inside of this compartment are painted flat black to prevent reflections.

The photocell is mounted on a small three-terminal strip. A single phone tip is soldered to the center terminal of this strip. To adjust the distance from the load lamp to the photocell, plug the phone tip into any of the four tip jacks as shown in the photo. The position and angle of the cell can be adjusted for the desired light pickup from the load bulb. The photocell is a Vactec cadmium sulphide unit and came from a Radio Shack No. 276-600 package which contains four cells (for only $1.19). However, there are many other similar cells that will work just as well. The other components are standard, and the schematic is self-explanatory.

The two variable resistors in the dc circuit of the load lamp are wire wound and should have at least two watts dissipation rating. All capacitors must be mica or ceramic suitable for high-frequency use. The variable resistor in the photocell circuit is an ordinary carbon control. The rf chokes are the ordinary pie-wound variety and are not critical. A value of 1 mH will work just as well, but the chokes must be of good quality for good vhf performance.

operation

Operation of the meter is simple. Select a bulb from table 1 with a power rating higher than the power that you plan to measure. The bulb must be operated below full brilliance for the best results. Just couple the output of the transmitter to J1 and adjust for maximum output (as indicated by the bulb) while maintaining the dc input to the transmitter at the desired level. At this point, if the bulb appears to be near full brilliance, replace it with a higher wattage bulb.

Turn the transmitter off and expose the photocell to room light. This can be either daylight or artificial light of about the same level as you would use to read

table 1. Pilot bulb characteristics.

No.	bead color	volts	amps	max power (watts)	resistance (ohms)
40A	Brown	6-8	.15	1.0	40
43	White	2.5	.50	1.25	5
44	Blue	6-8	.25	1.5	25
47	Brown	6-9	.15	1.0	40
49	Pink	2.0	.06	.12	33
49A	White	2.1	.12	.25	17
51	White	6-8	.20	1.2	33
53		14	.12	1.7	110

Resistance is for the power shown and will vary with a change of input current.

this magazine. Turn S1 on and set the meter M1 to full scale by adjusting R1.

Replace the light tight cover and turn the transmitter on. If the meter reads more than about 85% of full scale or less than 20%, change the distance or position of the photocell until the reading is somewhere between these two values. If the meter reads too low, an accurate reading cannot be made. If it reads too high, small changes cannot be seen.

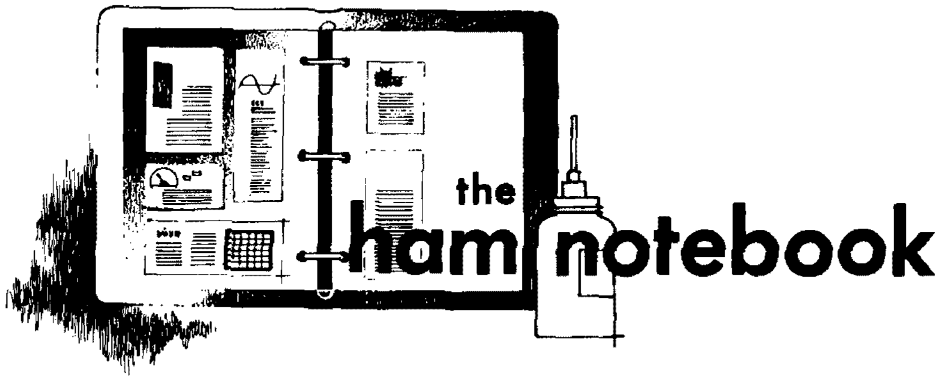
Note the meter reading and shut down the transmitter. With a battery or low voltage supply connected to *DC IN*, adjust R2 and R3 until M1 reads the same value as noted for the transmitter. The dc power computed from M2 and M3 will now be the power output of the transmitter.

A somewhat different method is used for very low power. Use a type 49 pilot lamp as a load and position the photocell as close to the lamp as possible for the best transfer of light. Adjust the dc lamp current to make the photocell meter read about mid scale. Note the dc power as read by M2 and M3. With the dc power still on, the rf power is applied to the bulb. The reading of M1 will increase. Reduce the dc input to the lamp until the meter M1 returns to its original reading (the reading before rf was applied). The difference between the first dc-input reading and the second dc-input reading is the amount of power removed to keep the filament level constant. This is equal to the amount of rf power that was added.

Many amateurs will feel that the photocell method of dc substitution is too long and involved for power measurement. They are right. However, it is just about the only accurate way to measure low power over such a wide frequency range and at a cost that the ordinary experimenter can afford. It is one of the very few methods that can be put in operation anyplace without using rf calibration standards.

Laboratory work has shown this method to be about as accurate or as sloppy as you care to make it.

ham radio



scr load protection

The unit described here uses a silicon-controlled rectifier with a biased gate as the heart of a circuit-breaker system which trips when *any* predetermined voltage has been reached. It can be used to trip at overloads or underloads. This is accomplished by using a biasing voltage between cathode and gate in connection with a second voltage which either adds to or subtracts from it. When adding voltages the device trips on overload; when subtracting, on underload.

R1 is a voltage divider and R2 a dropping resistor (fig. 1). It is the supply voltage minus the voltage between the anode side of R1 and the gate side of R2 that establishes the firing voltage of the SCR. If another voltage source is inserted between R2 and the gate, it can become the critical voltage in the firing of the SCR. This voltage can be introduced to reinforce or buck the voltage between the anode and the gate side of R2. This process determines whether the SCR fires on increasing or decreasing current in the monitored circuit. The voltage is developed across Rx and should be about one volt. As an example, take a transmitter whose plate current is not to rise above 150 mA. Then Rx = E/I = 1/.15 = 6.6 ohms.

R1 is adjusted so that the SCR *just* fires when 150 mA flows through Rx. Current through both the SCR and Rx is then turned off briefly, and R1 is backed off a degree or two. Current through the SCR and Rx is again turned on. The SCR should not fire. If the current through Rx has to be increased too much to fire the SCR, move the slider back toward the original firing position just a shade. A position should be found where the SCR does not fire until the maximum allowable current is exceeded.

If the unit is to be used to cut off current to a circuit when current in that circuit drops to some level, Rx is attached to buck the voltage between the anode and the gate side of R2. Again, adjust R1 after the proper current has been established through Rx. R1 is adjusted, as before, this time getting the SCR to fire when the current through Rx drops to the predetermined level.

When the monitored circuit is low voltage and low current (say a driving transistor in an audio or rf stage), a relatively high-impedance input dc amplifier (fig. 2) can be used to provide the voltage across what was called Rx earlier. An n-channel fet (MPF-103) and a germanium pnp transistor (HEP 3) do the job nicely. Rx here is a 300-ohm resistor which can be connected into the circuit

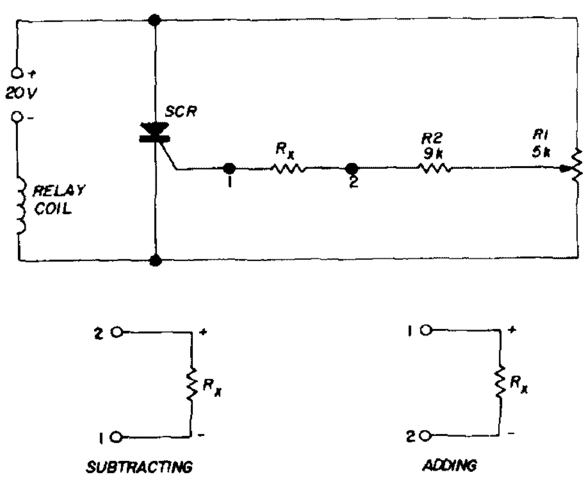


fig. 1. Basic SCR and relay circuit along with ways of inserting Rx.

of fig. 1 to reinforce or buck just as in the situation described earlier. The 20k variable resistor is adjusted (with nothing across the 5k variable resistor) to the point where current through Rx *just* reaches a minimum. Then the 5k resistor (with the load which it is monitoring connected across it) is adjusted to provide a reading of between one and three volts across Rx. The two rf chokes isolate the circuit from rf which might cause erratic triggering of the SCR.

The power supply should be capable of supplying the current and voltage necessary to actuate the relay, and the scr should be capable of handling these quantities, too. The power supply for the dc amplifier — if one is used — should be separate from the one for the relay. I use a model racing car power pack rated at 20 VA output, and a heavy-duty 24 V 6pdt relay with a 180-ohm coil. For the dc amplifier I use a small battery pack with eight AA cells in series. Since only three to eight mils are drawn, the battery should last quite a while.

George Hirshfield, W5OZF

sequential switching

This switching circuit provides delay in the make *and* break modes of switching and can be used to protect frontend transistors, diodes and coils due to momentary simultaneous operation of receiver and transmitter. The operation of

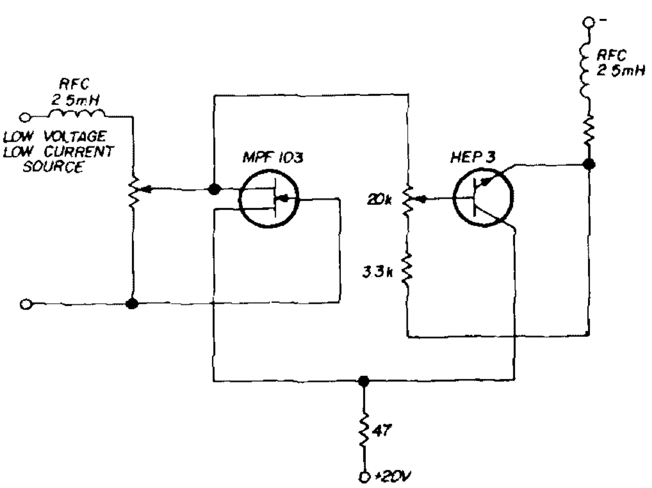


fig. 2. The dc amplifier.

the circuit is very simple, and the device is not at all delicate.

When S1 is thrown from receive to transmit, current flows to both the relay coil and the capacitor. The large capaci-

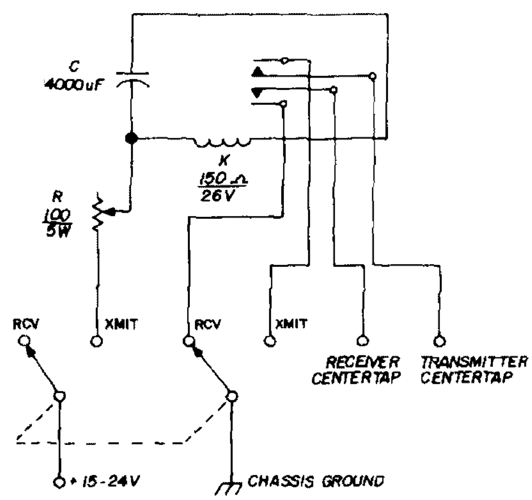


fig. 3. This circuit provides delay in make and brake switching.

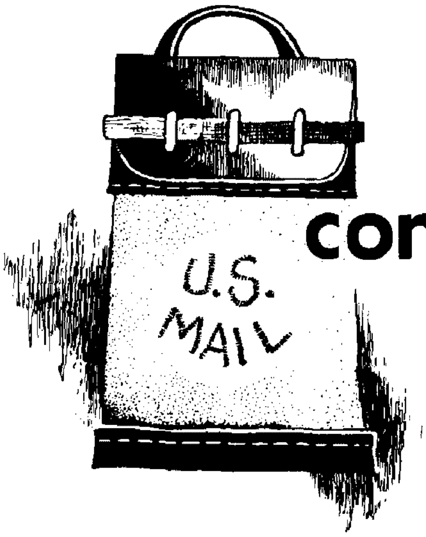
tance across the coil has the effect of acting as a near short circuit until, while charging through R, the energizing voltage for the relay is reached. When S1 is thrown from transmit to receive, the capacitor is discharged through the relay coil. Both processes occur, of course, in a finite and adjustable amount of time. The net effect here is about one-third of a second delay when switching from *either* mode to the other. The values shown are not critical. Voltages and resistances may vary according to relay requirements, but operation should be similar to what is described here.

The actual switching takes place when the centertaps of the receiver and transmitter power transformers are grounded through S1 *and* the contacts of the relay.

Contacts of the relay may be ganged for operation of additional circuits such as indicator lights, antenna switching or other relays for additional sequencing of other circuits.

The parts for this gimmick are generally available and can probably either be found in the junkbox or purchased from a single supplier.

George Hirshfield, W5OZF



comments

past is prologue

Dear HR:

I can't resist a few comments on your June editorial. The old timers you heard on 40 meters expressed the opinion that amateur radio today lacks the allure it used to have, that it has less uniqueness and fewer opportunities to offer. As a private pilot I hear the same song around airports. In part time marine radio work I hear much the same thing among boatmen.

I think the whole gang had better join a group I met last summer. I was invited to display a 66 year old car steam engine I've been saving to put in a boat. I met a gang of enthusiatic old timers who really get a kick out of their hobby. If amateurs feel jaded at the thought of competing in the technology race, pick up a hobby where the technology hasn't seen much change in the last 60 years, and isn't likely to change much in the next 60, either.

To me, the complaints of hobbyists who are caught up in the technological race are symptomatic of the state of the nation. There never has been an era when so many had so much yet were so dissatisfied. By comparison, I get a kick out of my wife's gardening adventures. Year in and year out she never tires of creating a riot of color from frost to frost. It is a tremendous source of satisfaction for her, and, perhaps more important, a chance to get away from the everyday rat race.

I think we are beginning to see a glimmer of that same thing in amateur radio today. The QRP gang are tossing aside all the usual equipment in order to create something different. They're having fun, but only because they've broken away from the crowd. Instead of saying, "Me, too," they are once again individualists. We need that sort of thing badly.

I've been in radio for 40 years, yet I wish I had 40 more years. The opportunities in amateur radio today are to me, incredible. I only hope that the QRP movement is only the start of a move towards greater individualism that will lead to new performance criteria, and once again, the opportunity for technical advances by the amateur body.

Your editorial suggests one such advance. "We at *Ham Radio* note that the day of the amateur-built receiver may have passed in favor of the vastly superior and less expensive commercial version." You are correct in this respect: The available receivers today could be compared to a luxurious car with power everything and loaded with accessories satisfactory to a majority of buyers, and perhaps to many a status symbol. Yet any one of maybe 100,000 kids could "homebrew" a performance car that could cream that Detroit iron on a drag strip.

Maybe we might see conditions in amateur radio reach the point where the individualists see the situation correctly. That the best receivers built today, while excellent, are not in truth "vastly superior" in performance. In fact, many of the most expensive receivers are indeed inferior.

A rabid DX man could, for instance,

settle for a single band receiver – maybe 20 meters. Many fine commercial receivers have sensitivity greater than you can use, due to the limitations of atmospheric noise level. However, over the span of frequencies the 20-meter band represents, noise distribution will be relatively uniform. If so, suppose he narrows the front end passband to half that of the commercial receiver? In a given situation, his receiver will show a 3-dB improvement in signal to noise ratio. If he cuts that bandwidth again in half, the improvement is 6 dB. If he can cut the front end passband by a factor of 8 he is 9 dB better. Of course, at this point the receiver front end may have to be re-peaked for relatively minor changes in frequency. Already his cross-modulation capability is improving. Careful selection of rf and mixer will enable him to top any commercial receiver in this respect. It would take some care in determining optimum *gain* distribution, too. In the same manner he can top the best receivers in stablity and selectivity, and still not have as much invested in parts cost.

If a trend developed in the direction of *true "competition" receivers*, I wouldn't be surprised to find a parallel to today's automotive performance parts business. It wouldn't be big business any more than the speed shops are in comparison to Detroit. Such a trend would sure separate the men from the boys on the amateur bands, particularly if as much care were devoted to antenna design.

That's one example, and there are a lot more loopholes in amateur radio that enterprising individuals can plug to their own advantage. I just hope we are seeing the start of a trend.

Bill Wildenhein, W8YFB

ground-plane antenna footnote

Dear HR:

I was intrigued by the simple elegance of Peter Brekken's ground-plane antenna in the May, 1972, issue and built the single-band, 20-meter version. I had a reasonable duplicate of the vertical element already mounted on my roof – a 40-foot heavy-duty tv mast that had supported a variety of wire antennas. Brekken's design calls for 75-ohm coax for the transmission line; however, I wanted to use RG-8/U 50-ohm line, which I had on hand. I also wanted to use existing materials and junkbox parts to keep expense at a minimum. The total cash outlay for my version of the antenna was $1.85. The following comments are offered for those who might want to try this excellent antenna with the more common 50-ohm coax cable.

The feed point of my antenna is almost exactly ¼ wavelength above ground at the design frequency (14.15 MHz). I built a tuning unit based on the formula in Brekken's article, but added a few extra turns to the coil. The entire arrangement was made of odds and ends available around the house – nothing sophisticated – just ordinary hardware and the usual junkbox parts.

For example, the coil was made from tv "ground wire," which is no. 8 soft-drawn aluminum wire. This material is self-supporting if a reasonable form factor *is used for a* coil. My coil is 3 inches in diameter by 6 inches long, supported at both ends by standoff insulators. A short length of RG-8/U coax connected between an appropriate tap on the coil and the antenna vertical element adds support to the coil. The coil, which is about 3 μH, was *wound over a* quart beer bottle (empty) and hand formed. A weather shield covers the entire assembly.

The impedance characteristics of my 5/8λ vertical antenna were measured with a General Radio model 916A bridge. Check points are:

Frequency (MHz)	resistance (ohms)	reactance (ohms)
14.0	128	-j305
14.05	118	-j223
14.15	108	-j215
14.20	92	-j196
14.25	82	-j175
14.30	74	-j158

The capacitive reactance of -j215 ohms at 14.15 MHz was tuned out by making many trips between the transmitter and the coil at the antenna base until a minimum swr was achieved by adjusting the coil tap. The lowest swr obtainable was 2.5. This meant that the -jX component of impedance was accounted for, but the resistive component, 108 ohms, still had to be compensated to achieve a good match at the antenna resonant frequency.

A ¼λ transformer made of RG-8/U cable connected in series with the transmission line and antenna feed point solved this problem. The ¼λ transformer was made according to

$$Z_t = \sqrt{Z_a Z_o}$$

where

Z_t = characteristic impedance of transformer (ohms)
Z_a = antenna resistance (ohms)
Z_o = characteristic impedance of transmission line (ohms)

Substituting values,

$$Z_t = \sqrt{(108)\ (52)} = \sqrt{5610} = 75$$

A 17-foot length of RG-11/U cable was used to make the matching transformer. This was purchased from a local surplus house for $1.00. With the ¼λ transformer installed between the 5/8λ vertical and the RG-8/U transmission line, the swr between 14.0 and 14.3 MHz was pretty low; the mean value was measured at 1.15.

A simple weather shield completes the assembly. I found a plastic tub in a local supermarket for 85c, which I mounted over the tuning assembly at the base of the antenna.

If you don't have the exact materials on hand that are called for in published articles, try substitutes. That is what makes ham radio fun.

Alf Wilson, W6NIF
San Diego, California

microwave equipment

Dear HR:

I wish to point out that if you are willing to search out suppliers and take your time, the equipment listed in the article by W. T. Roubal in your June, 1972, issue on "Getting Started in Microwaves" can be purchased for 10% or less of the costs listed in his article.

I have many 2K25 formerly 723A/B Klystrons that I have purchased for 50c to $1.00 (Do not tune the cavity too often as the bellows will fatigue and let air into the tube.)

I have several variable attenuators bought for less than $5.60.

The only problem with this equipment is that it has not been checked out for its working condition. This means that you must do the checking yourself. The procedures for doing the checking can be found in the Berkley Lab series books A, B and C or, for a much deeper explanation, the MIT Radiation Laboratory series now put out by Doer Press for 2 to 4 dollars each.

The procedures are fairly simple if you ignore the math in these books. Many tests can be run with nothing more than a 2K25, its power supply and waveguide mount, a waveguide mounted detector and a meter to measure detector current.

If you want to make swept frequency measurements you must have an oscilloscope that has at least a dc to 1-kHz bandwidth.

The slide screw tuner of fig. 5 in Roubal's article can be made into a slotted line by replacing the micrometer with a detector diode and a small antenna that sticks into the guide like the plunger of the micrometer.

This setup will allow you to measure standing waves in the wave guide. For an sase I am willing to answer any questions on the above subject.

Edward A. Benjamin, WA1HYX/4
1010 13th Street North
St. Petersburg, Florida 33705

prologue to the future

Dear HR:

It has been my observation that it is *participation* in the act of communication, rather than the material communicated, that has been the unique attraction of amateur radio over the decades. I think your editorial in the June issue strongly (and correctly) points up the attrition of this aspect of the game, but I cannot share your optimism.

We all seem to agree that, whether we like it or not, amateur radio is losing its unique flavor. In fact, it is tending to become just another communications system, operating roughly in parallel with Ma Bell and her sisters.

It is the intimate contact with the inner techniques and difficulties that enthralls most of us who read *ham radio*. We enjoy building gear, however simple, tuning it up, and solving real problems in the communications process. For a large number of us this may even be the whole game.

After we've gotten things working as we think they should, it's time to tackle another problem. What gets communicated via the system is, for the most part, incidental. Unfortunately, the rapidly advancing techniques place most of us further and further from as many significant problems as were formerly within our reach – day by day we're being phased out.

It is my personal belief that no sophistication of technique nor improvement of communications efficiency can ever, in any way, replace genuine human satisfaction. The proven pleasures of personal involvement, even of the simplest sort, are irreplaceable by mere machinery, no matter how sophisticated. When Mr. Walker's *vaunted satellite is up there, it* will be fun to "work through" a few times, just to satisfy ourselves, but then what?

Just as the finest product of American professional engineering cannot truly replace the beloved haywire in the true home-brewer's heart, so the finest satellite will never, for many of us, replace the inefficient, capricious ionosphere. For the satellite is only machinery, while the ionosphere, like the haywire, is *adventure.* And adventure is the priceless ingredient in amateur radio.

C.F. Rockey, W9SCH
Deerfield, Illinois

tape head cleaning

Dear HR:

The Multiple Audio Distribution (MAD) System at Western Michigan University consists of Magnecord 1048 tape machines which are in use almost continuously. Currently, seventeen machines are in use, with eleven machines in use 24 hours a day, 7 days a week, 48 weeks a year. Some of the machines are six years old, while the newest is three years old.

The machines are cleaned Monday through Friday with alcohol and xylene. Xylene is used on the tape heads, *metal* tape guides, stabilizer rollers and capstan. The xylene, as mentioned in your May issue, requires care in use as it is damaging to plastics and some paints. The ability to dissolve the binder in the tape oxide is what makes the xylene worth the care needed in use. A Q-tip full of xylene will remove the biggest glob of dirt and oxide.

Alcohol is used on the pinch wheel, because, when it is used often, it is sufficiently strong to dissolve the oxide without drying out the pinch wheel. The staff at WMU-TV uses xylene on their video tape recorders, but in an emergency when a head clogs during playback they give it a shot of Freon TF* which loosens the clog without dissolving the oxide on the tape as xylene would do.

James R. Buchanan
Western Michigan University

*Freon TF is manufactured by the Tex-Wipe Company, Hillsdale, New Jersey.

new products

transistor curve generator

A new easy-to-use transistor curve generator, known as the Model TCG-1, is now available from Caringella Electronics, Inc. This high performance instrument is designed for experimenters, hobbyists, schools, engineering labs and service shops. Transistors and other semiconductor devices can be tested in or out of circuit with this versatile new test instrument. The transistor curve generator is used with any oscilloscope, and displays the dynamic characteristics of both npn and pnp transistors, fets, mosfets and dual-gate mosfets, diodes, zener diodes, tunnel diodes and other devices.

The instrument incorporates all the circuits required to generate the base steps and collector sweeps. The collector sweep generator provides a ± 10-volt saw-tooth, operating at a frequency of 550 Hz, for a flicker-free display. A fully regulated power supply, utilizing a ± 15-volt IC regulator, and a solid-state LED panel indicator light, are also features of the solid-state design. Operation is simple and straightforward due to the minimum number of front panel controls. Several unique features highlight the Model TCG-1: direct transistor "beta" readout is provided on the front panel base-drive control, npn and pnp transistors can be tested consecutively without changing controls or switches and vertical and horizontal channels of the oscilloscope are calibrated simultaneously for accurate readings.

In addition to displaying the collector current versus collector voltage family of curves, the instrument also tests transistor "open base" and "shorted base" collector breakdown voltage. Two TO-5 transistor test sockets and a set of three binding posts are provided on the front panel and are selected by a 3-position lever switch.

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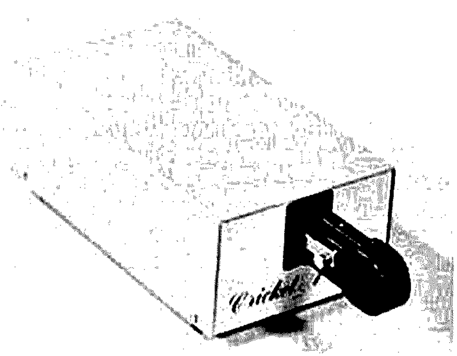
Semiconductor devices can be compared or matched easily, and even unknown transistors types can be identified quickly. The versatile binding posts will accommodate external test leads or a variety of test fixtures for production testing. A set of three color-coded test leads is provided with each instrument.

The Model TCG-1 is available in kit form, complete with all parts, high quality glass-epoxy printed circuit board, wire, solder and step-by-step assembly instructions. Kit price is only $79.95 complete. Factory wired and tested units are only $99.95 complete. All shipments are F.O.B. Upland, California, and the kits and wired units are available for immediate delivery from stock.

A free data sheet, complete with technical specifications, schematic diagram and circuit description, is available on request. For further information contact: Caringella Electronics, Inc., Box 327, Upland, California 91786 or use *check-off* on page 110.

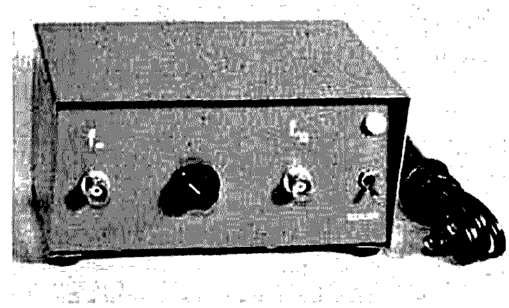
cricket keyer



Data Engineering has introduced a new small keyer with a built-in key, sidetone oscillator, speaker and ac power supply. The key features jam-proof spacing, self-completing dots and dashes, a heavy duty 300V, 1A relay, speed range from 3.5 to 50 wpm and a jack for powering the unit directly from 12 Vdc.

Ready-built and with Data Engineerings five-year guarantee, the unit sells for $49.94. For more information write to Data Engineering, Box 1245, Springfield, Virginia 22151 or use *check-off* on page 110.

uhf digital frequency scalers



A ten-fold increase in the range of many frequency counters is possible with the use of Belmont Spectrum Research frequency scalers.

Two Models are offered. The *Model A* includes a two-stage preamplifier and is the more sensitive. Typically, it requires inputs of less than 3 mV up to 100 MHz, 15 mV at 200MHz, 45 mV at 250 MHz, 75 mV at 275 MHz and 120 mV at 300 MHz. Guaranteed to 280 MHz. $90 plus California sales tax and shipping charge of $1.50.

The Model B has no preamplifier and thus requires greater signal input. Typically, it requires 20 mV up to 100 MHz, 140 mV at 200 MHz, 260 mV at 250 MHz, 340 mV at 275 MHz and 800 mV at 300 MHz. Guaranteed to 260 MHz. $70 plus California sales tax and shipping charge of $1.50.

These scalers use Fairchild's 95H90, high-speed, ECL, MSI devices for the basic divide-by-ten function. Each model also uses Fairchild's 7805 regulator, thus assuring a highly reliable and constant 5.0-volt power supply. They were first introduced at the 1972 West Coast VHF/UHF conference in May.

For more information, contact Belmont Spectrum Research, 1709 Notre Dame Avenue, Belmont, California 94002, or use *check-off* on page 110.

proximity detectors and metal locators

Written in a comprehensive but easy-to-read style, the updated second edition of "How To Build Proximity Detectors and Metal Locators" by John Peter

Shields, contains a reservoir of valuable information on the principles and circuits used in proximity detectors and metal locators. It also covers the closely related theremin, the source of many of the eerie sound effects heard on the radio, on television and in the movies.

This handy guide begins with a simple explanation on the basic types and functions of proximity detectors and metal locators, and elementary proximity detector and metal locator projects are given. Here any novice or experienced electronics enthusiast can learn how to build basic one or two stage circuits. Then the more complicated or advanced electronics projects containing additional stages or more complex circuitry are included. Examples of two advanced circuits analyzed are the Hall-effect metal detector and the fm-discriminator proximity detector.

These circuits are used for a wide variety of purposes including burglar alarms, touchswitches for activating lights and other electric devices, locators for finding pipes and studs in walls and floors and for locating lost metal objects such as coins, jewelry, keys and tools in the ground or under water. In addition, this edition has important facts on several new circuits, some of which utilize a modern development in solid-state components — the silicon triac thyristor.

Each circuit has been thoroughly tested and uses components that are available at most electronic parts distributors. Complete parts lists and illustrated assembly instructions are included for each project reviewed.

Students, technicians, hobbyists or anyone else interested in learning about or building proximity detectors and metal locators will enjoy this "do-it-yourself" book.

This 160-page softbound book is available from Comtec Books, Greenville, New Hampshire 03048 for $3.95.

abc's of electronics

Howard Sams has introduced a second edition of "ABC's of Electronics" by Earl

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J. Waters. This new edition is an easy-to-grasp, but comprehensive introduction to the broad field of electronics. The author avoids complicated technical concepts and mathematical terms as much as possible and relies on simple language and analogies familiar to everyone.

The text presents a detailed analysis of the principles of electricity, functions of atoms and electrons, magnetism and solid-state physics. Individual chapters are devoted to electrical resistance, capacitance and inductance. The remainder of the book deals with alternating currents, circuit impedances, electromagnetic radiation, vacuum tubes, transistors, integrated circuits, radio wave production and propagations and the various electro-mechanical devices.

Each chapter concludes with a number of review questions; the corresponding answers are located in an appendix. The appendices also contain valuable reference data on electronic standards, mathematical formulas and color codes of resistors and capacitors.

The new edition is a reservoir of knowledge supplemented by many illustrations; it was written to keep up with the rapidly changing field of electronics.

The 160-page, softbound book is available from Comtec Books, Greenville, New Hampshire 03048 for $3.95.

crystal filter

Housed entirely in a standard HC-6/U can, the new SM 107S04 crystal filter offers a four-pole crystal filter centered on 10.7 MHz. The new unit features a minimum ±7 kHz bandwidth at -40 dB, ±12 kHz maximum bandwidth at -40 dB and a maximum 3 dB insertion loss. Maximum ripple is 1 dB, ultimate attenuation is 60 dB minimum. The attractive packaging of the filter uses one of the pigtail leads for input, one for output and the case is grounded. Units are available from stock postpaid for $15.95. More information is available from Spectrum International, Box 87, Topsfield, Massachusetts 01983 or by using *check-off* on page 110.

Hallicrafters catalog

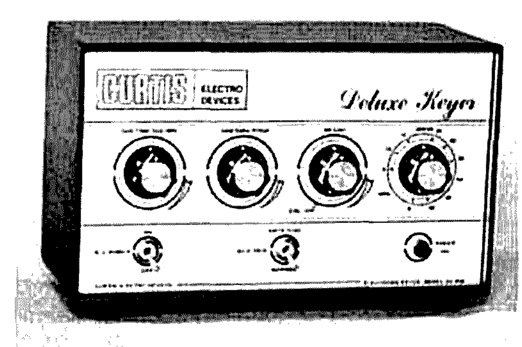
The Hallicrafters Company has issued a four-page, well-illustrated, short-form catalog "You should be talking with a Hallicrafters." It features the company's entire line of shortwave receivers and amateur radio equipment.

Designed for both the beginner and the experienced amateur, the easy-to-read catalog gives the major features of the SR-2000 "Hurricane" transceiver, HA-20 remote vfo/swr console, HA-1A T.O. keyer, SR-400A "Cyclone III" transceiver plus the entire line of accessory equipment.

Also featured in the catalog are the company's SX-122A communications receiver and the SX-133 high-performance receiver. The latest addition to the Hallicrafters line, the FPM-300 solid-state hf ssb/CW transceiver, is also included in the catalog as well as in a separate data sheet.

For a copy of "You should be talking with a Hallicrafters" or the FPM-300 data sheet, write The Hallicrafters Company, Department PR-300, 600 Hicks Road, Rolling Meadows, Illinois 60008 or use *check-off* on page 110.

iambic keyer



An iambic keyer with many new convenience features has been announced by Curtis Electro Devices. The EK-404 is an offshoot of the EK-402 programmable keyer introduced about a year ago and employs the same styling and features with the exception of the message memory.

Standard features in the EK-404 in-

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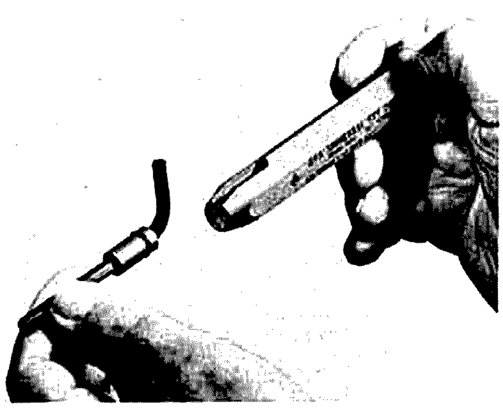
clude jam-proof dots, dashes and spaces; iambic or standard operation; dot memory; variable weighting; tune switch; built-in sidetone and speaker; a self-contained 115-Vac power supply and complete rf immunity. New features include a connection for 12-Vdc mobile or portable operation, front panel control of sidetone pitch, a self-test mode, two sidetone outputs (Hi-Z, Lo-Z) and provision for either grid block and cathode keyed or solid state rigs. A manual key jack is also provided.

Price of the EK-404 is $124.95 complete with all cables and connectors. It is available direct from the factory or from dealers. For further information write Curtis Electro Devices, Box 4090, Mountain View, California 94040 or use *check-off* on page 110.

universal key driver



A new hand tool for driving L-keys has been introduced by Jensen Tools and Alloys. Known as the GLA Universal Key Driver, this new tool will be found particularly useful to anyone who uses special screws in his hobby or vocation.

The GLA tool is a common sense, high-torque driver for any type of English or metric screw key up to 0.217 inch (5.5 mm). It accomodates hex (Allen), spline (Bristol), clutch-head, Scrulox, cross recessed (Phillips), Reed and Prince, or any other type of L-key within its dimensional capability. The key is simply slipped into one of nine different bushings (any one which clears), the bushing is slid into the handle and the tool is ready for use. There are no set screws to tighten and no broached holes or plastic to strip or break.

Torque is tremendously increased over that of conventional driver designs, being limited only by the ultimate strength of the key being driven (over 30 foot-pounds). The tool may be used as a "T" driver by applying the short arm of the key to the screw and rotating the handle in propeller fashion instead of axially. Also, a socket wrench can be applied to the handle hex, to obtain any desired reach. The GLA tool is 5 inches long by 3/4-inch hex and is made of hard aluminum. Standard O-rings keep the bushings in place. Bushings have graded hole sizes to accept all keys of any cross section type, up through 3/16-inch hex.

A twenty-piece set is offered which includes the basic GLA tool, nine bushings, a nine-piece hex-key set including all sizes from 0.050 to 3/16 inch and a wooden box. The wood box serves as an island stand. The set sells for $14.25 postpaid. For further information, contact Jensen Tools and Alloys, 4117 North 44th Street, Phoenix, Arizona 85018 or use *check-off* on page 110.

HEP data sheets

Engineering and design data sheets are now available for many Motorola HEP semiconductor products. These descriptive sheets contain complete and comprehensive information on the specified devices including design curves, rating charts and application schematics. A Motorola spokesman explained, "We believe that we are the only manufacturer specializing in this type of sales, to make information this comprehensive available to the hobbyist and radio amateur. These data sheets will help eliminate guesswork and 'make do' applications."

Copies of the data sheets and additional information are available from HEP representatives throughout the country. HEP is Motorola's sales program for making semiconductor devices readily available to the hobbyist-experimenter and to professional service dealers through a nationwide network of authorized suppliers. Motorola HEP Semiconductors, Box 20924, Phoenix, Arizona 85001.

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breadboards

EL Instruments has a very comprehensive line of electronic breadboards, semiconductors and experimental power supplies. Features of the electronic breadboards include solderless connections and adaptability to all standard electronic components including direct plug-in of DIP ICs. Interconnections are made either by internal ties or externally with ordinary hookup wire — no special cords are needed. The units are completely reusable and the nickel-silver contacts are designed for over 10,000 component insertions.

Boards come in many different forms including plug-in boards for card racks, a standard screw-mounted plain breadboard and a number of deluxe breadboxes which include boards and cabinets for more complex or more permanent circuits.

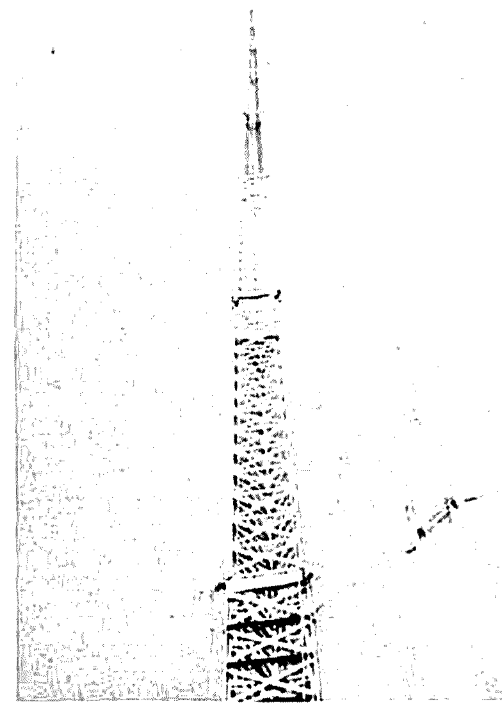
The complete catalog gives more details on all these units and on other experimenter's supplies. The catalog is free from EL Instruments, Inc., 61 First Street, Derby, Connecticut 06418 or by using *check-off* on page 110.

antenna catalog

A comprehensive, new 96-page general catalog listing over 250 models of professional communications antennas has been released by The Antenna Specialists Company. Complete mechanical and electrical specifications and radiation patterns are provided, along with full details of mounting options. The catalog covers full lines for all land-mobile antennas, plus selected base and mobile antennas for professional monitoring, marine, avionics and CB.

Of particular interest to amateurs, there is plenty of general information on transmission line characteristics, side-mounting patterns and element cutting charts. The catalog is available on request to Professional Communications Department, The Antenna Specialists Company, 12435 Euclid Avenue, Cleveland, Ohio 44106 or by using *check-off* on page 110.

atmospheric probe



Tri-Ex Tower Corporation has completed field tests on their new 205 foot XM-205 "Atmospheric-Probe" crank-up tower. This new tower has been under development for several months and was designed, engineered and manufactured by Tri-Ex.

The three lower sections of the new tower are X braced and the upper five sections are M braced. The tower has a total standing weight of 2500 pounds. At the start of erection, the tower stands 40-feet high with a triangular base of 27¾ inches and reaches a 205-foot full erection height with 9¾ inch triangular top section.

The new "Atmospheric-Probe" tower features "safety lock fixtures" at every one of the seven levels and is raised approximately three feet per minute by a self-contained motorized winch.

This tower's stiffness and strength is enhanced by the use of a generous five-foot overlap between sections. The tower is progressively guyed, as the sections extend upwards.

This tower, to date, is the largest and tallest crank-up tower manufactured by Tri-Ex. Other models in the XM series

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will soon be available.

Complete information is available by contacting Tri-Ex Tower Corporation, 7182 Rasmussen Avenue, Visalia, California 93277 or by using *check-off* on page 110.

scr manual

General Electric has produced a new fifth edition of its "SCR Manual." This latest edition, issued on the fifteenth anniversary of General Electric's invention of the SCR, consists of over 75% new text and covers SCRs, TRIACs, uninjunctions and triggers.

An idea of the comprehensiveness of this new book can be gathered from this list of the *major* chapters: Construction and Basic Theory of Operation; Symbols and Terminology; Ratings and Characteristics of Thyristors; Gate Trigger Characteristics, Ratings and Methods; Dynamic Characteristics of SCRs; Series and Parallel Operation; The Triac; Static Switching Circuits; AC Phase Control; Motor Control Employing Phase Control; Zero Voltage Switching; Choppers, Inverters and Cycloconverters; Solid State Temperature and Air Conditioning Control; Light Activated Thyristor Applications; Protecting the Thyristor Against Overloads and Faults; Voltage Transients in Thyristor Circuits; Radio Frequency Interference and Interaction of Thyristors; Mounting and Cooling the Power Semiconductor; SCR Reliability; Test Circuits for Thyristors; Selecting the Proper Thyristor and Checking the Completed Circuit Design. There is a final chapter on garnering specific device application notes.

Copies of the book can be obtained from any authorized General Electric Component distributor or by sending $3.00 plus applicable tax to General Electric Company, SCR Manual, Department B, 3600 North Milwaukee Avenue, Chicago, Illinois 60641.

new HEP catalog

Approximately 38,000 semiconductor devices are cross-referenced to HEP replacements in the new 1972 Motorola HEP Semiconductor Cross-Reference Guide and Catalog. Included in the catalog are 1N, 2N, 3N, JEDEC, manufacturers' regular and special *house* numbers as well as many international devices, with particular emphasis on Japanese types.

A total of 467 different HEP items are included in this guide, including hardware, accessories, technical manuals and hobby project books. As in previous editions, the Motorola HEP devices are listed by type number with a packaging index, device dimension drawings and selection guide information.

This cross-reference guide and catalog is available free at local Motorola HEP suppliers throughout the country. It should be of particular interest to the electronic hobbyist and radio experimenter since it gives the minimum/maximum ratings and the electrical characteristics for the HEP devices as well as cross-reference information.

W9IOP's second op

The new, 6th edition of W9IOP's famous and useful Second Op is available now. This 10½-inch circular operating aid is covered with up-to-date information for the active DXer. The user sets the pointer to the call-letter prefix, and the Second Op displays the country name, WAZ zone number, continent and great circle beam headings from four different United States population centers. Also shown is the local-time to tocal-time conversion factor between the DX location and three United States time zones. Postage information for letters and QSL cards by air and sea mail along with the number of International Reply Coupons necessary for an airmail letter reply is also displayed. There are two boxes next to every prefix for recording the prefix worked and the prefix confirmed. Printed on the Second Op are complete instructions for its use and a list of North

American and worldwide QSL bureaus.

Previous editions of the Second Op have earned it a reputation as a valuable operating aid to identify, work and QSL DX stations. The operating aid is printed in three colors on sturdy card stock. Produced by Publications in Electronics, it is available for $2.00 from Comtec Books, Greenville, New Hampshire 03048.

pocket receiver

Old ideas seem to come back in many new forms. Induction wireless has reappeared in the Lowcom Systems wireless induction pocket receiver. The receiver, a tiny and attractively packaged unit, is designed to allow you to monitor a receiver, audio system, radio or TV without being confined to the immediate vicinity of the unit. The audio output of the receiver (or other system) is fed into a wire induction system. The pocket-sized receiver is inductively coupled to this output loop and feeds its output through an amplifier to a personal earpiece. This system allows complete freedom of movement while monitoring a receiver *without* bothering anyone else with a blaring speaker.

This particular system is patented, confirming one's suspicion that this particular little device is a spin-off of Lowcom's industrial paging and audio distribution systems. The unit is attractively packaged and comes with a brass plate neatly engraved with your call letters. For those who can remenber experimenting with large coils of wire, dry cells and old telephone microphones; it is quite a sight to see some of the old schemes existing in this modern, tiny, commercial box.

The unit sells for $24.95, postpaid with batteries, case, earphone, installation instructions and engraved call plate. Quantity discounts are available.

For more information write to Lowcom Systems, Inc., 10727 Indian Head Industrial Boulevard, Saint Louis, Missouri 63132 or use *check-off* on page 110.

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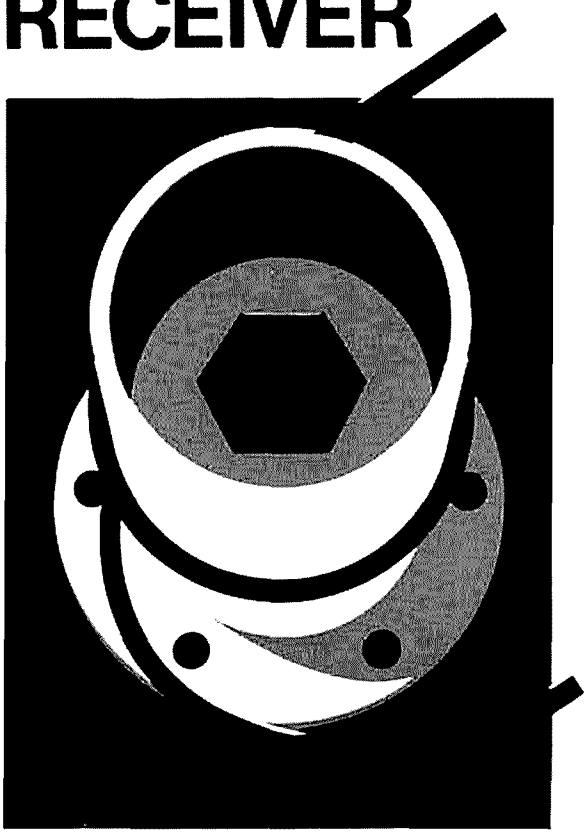
ham radio magazine

NOVEMBER 1972

VHF FM RECEIVER



this month

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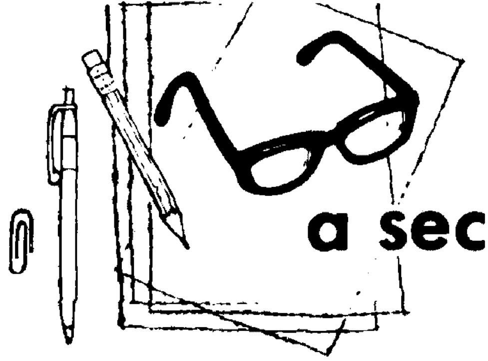


a second look

by jim fisk

The FCC has just approved some additional telephony frequencies within the amateur 80- and 40-meter bands. The Commission decided, however, not to permit any phone expansion on 20 meters, saying it would result in serious degradation to non-voice communications. The Commission also found that the 10- and 15-meter bands were not crowded enough to warrant the expansion of phone segments on those bands. The new phone segments become effective on November 22nd, 1972.

If you will remember, the FCC originally proposed to expand the telephony sub-allocations of the five amateur high-frequency bands in early 1971. Also proposed were changes to expand the operating frequencies for General and Conditional class licensees, and to provide additional phone sub-bands for Extra and Advanced class amateurs where they do not presently exist.

Numerous comments on the proposed rule changes were received from amateurs in the United States, as well as from foreign countries. A delegation from one country even made an official visit to the Commission to discuss the proposals. As can be imagined, foreign amateurs were strongly opposed to any telephony expansion in the United States, particularly on 20 meters.

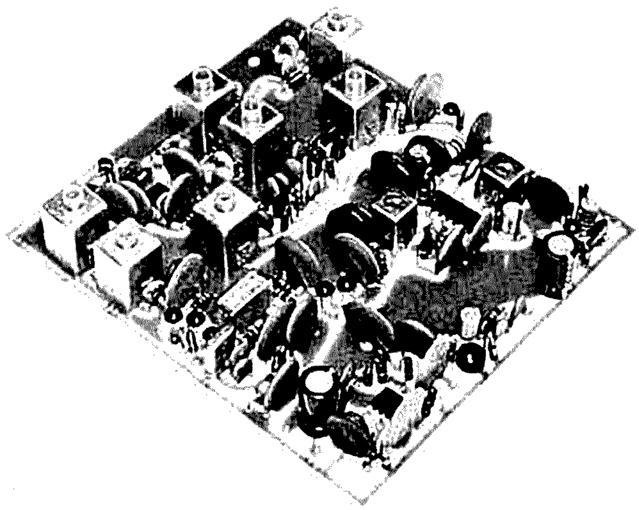
Since there are no formal, internationally agreed, sub-allocation plans reserving portions of any of the high-frequency amateur bands for any one type of emission, amateurs in various parts of the world operate under informal gentlemen's agreements. Therefore, many overseas amateurs predicted, if U.S. amateurs were *permitted to* expand *their* phone operation into the sub-band between 14.1 and 14.2 MHz, that non-U.S. phone stations *would retaliate* by moving into the 14.0 to 14.1 MHz segment, causing a deterioration in CW and RTTY communications. Hence, the Commission decided not to expand our phone privileges on 20 meters.

The Commission decided not to authorize an expansion of 50 kHz in the 75-meter band for phone operation by Amateur Extra and Advanced class licensees, as originally proposed, but to adopt an expansion of 25 kHz, 3775 to 3800 kHz, limited to Amateur Extra class. Advanced-class operators may now operate from 3800 to 3890 kHz, and General class licensees have phone privileges from 3890 to 4000 kHz.

The 40-meter phone segment has been expanded, as originally proposed, permitting phone operation between 7150 and 7200 kHz. As a result, the 40-meter Novice band has been relocated from 7100 to 7150 kHz. Amateur Extra and Advanced licensees may now use phone on 7150 to 7225 kHz, and the General-class phone segment runs from 7225 to 7300 kHz. The Commission did not adopt a proposal to provide a phone sub-band below 7100 kHz for contacts with stations in Regions 1 and 3, but *did* adopt the proposal to permit phone operation between 7075 and 7100 kHz for American stations located *outside* of Region 2.

To help beginning amateurs gain experience in CW operation, a new Novice segment was adopted for 28100 to 28200 kHz. Lightly occupied Novice segments 21.2 to 21.25 MHz and 145 to 147 MHz were deleted, leaving a 15-meter Novice band that runs from 21100 to 21200 kHz. And, big news for Novices, transmitters no longer must be crystal controlled.

Jim Fisk, W1DTY
editor



vhf fm receiver

A versatile high-band or low-band state-of-the-art vhf fm receiver

Jerry Vogt, WA2GCF, 182 Belmont Road, Rochester, New York 14612

Are you looking for a compact, low cost receiver to use with your new homebrew fm transmitter to make a good transceiver? Are you interested in trying fm without investing a lot of money for a transceiver right away? Do you need an extra receiver around the shack to monitor your local repeater or calling channel while you're off on another frequency? Do you and your wife have a private frequency for house-to-car that you would like to monitor? Would you like to build a low cost, but high quality, receiver for the commercial, marine or public safety band? Do you hate to buy something that you can build yourself, either for satisfaction or to learn what makes it tick? I might have the answer if you enjoy homebrew work! This can be an exciting winter project for you.

For several years, as I have developed numerous homebrew receivers, I had the feeling that more hams would be rolling their own if someone who had refined the art of receiver construction would show them the tricks or perhaps make a basic kit available. I had written many notes to kit manufacturers, trying to prompt them to put out module kits; thereby allowing the ham to build the difficult part of a

receiver from available kits and then custom finish the receiver by adding case, speaker, controls and other peripheral goodies. Evidently, the low market volume or lack of faith in the ability of the advanced ham builder caused the idea to be scrapped. Too bad!

Then, along came fm activity amongst hamdom. I got bit, and I jumped in head first as I usually do in any new venture. I turned all my homebrew efforts into building receivers (my favorite) for two-meter fm. At the same time, I began to go all solid state. The rapid growth of integrated circuits made receiver construction more fun by injecting a bit more magic into the art.

Over a period of three years, I developed nine fm receivers, including narrow-band receivers for two meters and the high public-safety band (I am also a volunteer fireman) and one for fm music (what the ICs were really intended for). The subject of this article is a receiver developed especially for readers of *Ham Radio*, the cream of the homebrew ham crowd. Unlike previous designs, this one is not tailored to my individual needs, but is designed to allow maximum flexibility consistent with low cost and ease of construction and alignment.

This is a big undertaking, since I vowed to make all the special parts available to allow anyone to build it. However, I am encouraged that it will be a big hit, because I have previously done the same thing with six- and two-meter preamps and other projects.[1] I was delighted with the response; especially the letters which followed from hams all over the world, indicating that hams do want to build, that homebrewing is not a lost art, and that all the average ham wants is a little headstart in knowing where to start and how to get parts.*

design considerations

Let me say from the outset that this design uses some of the fine goodies available to make the set reproducible, including mosfets. Anyone having read my earlier article will undoubtedly question my use of mosfets after having spoken out against them in the previous article. I am a big cascode pusher, being all in favor of its obviating the need for neutralization. The only real objection to mosfets was in their fragility, being susceptible to damage from static discharges before being soldered into the circuit. Recently, protected gate mosfets have been developed to eliminate that problem. The MPF-121 dual-gate mosfet used in this design is a homebrewer's dream; it is the answer to a "cascoder's" prayer. Besides being inexpensive, it, like other cascode devices, requires no neutralization. The noise figure is relatively low, and it is relatively immune to cross-mod overload.

features

Enough editorializing! From here on, I will discuss the circuitry and construction of the receiver. I will leave the finishing touches, such as case, controls, etc., to your imagination, and I will concentrate on the inner workings. I am sure that the average *Ham Radio* reader is above being sucked into building a project because of

*The following components are being made available in conjunction with this project. Be sure that you specify exactly what you want, including frequency bands. Complete parts kit for A1 assembly is available with undrilled circuit board for $54.95, or with pre-drilled board for $59.95. Six-channel oscillator board A2 parts kit is available with undrilled circuit board for $9.95, and with pre-drilled circuit board for $12.95.

Channel crystals are available separately. ZIP certificate for monitor crystal (shipped directly from the factory when you send the certificate) is available for $5.00. A kit containing a 3 x 5-inch, 16-ohm oval speaker and squelch and volume controls is offered as an accessory at $5.95.

Delivery on all items is subject to availability at the time of receiving your order. Every effort will be made to supply your requests as soon as possible, and you will be notified if there will be a delay in shipment. Expected time of delivery is about two weeks. Quantity prices are available on request for clubs and individuals interested in local distribution to hamfests, etc.

All prices are in *US$* and include postage *for* domestic parcel post. Write to Hamtronics, Inc., 182 Belmont Road, Rochester, New York 14612.

a fancy panel which he probably doesn't have equipment to reproduce. Many articles have been written around finishing techniques if you are interested, and the variety of methods is endless.

The basic receiver consists of one printed-circuit board as pictured. The board, dubbed assembly A1, incorporates a one-channel oscillator, with a separate (A2) board available to allow six-channel operation, if desired. The advantage of this scheme is that if a user wishes to monitor only one channel, small size and low cost makes such an application attractive. However, for multichannel operation, it *is still* very easy to add the extra channels. One recommended procedure is to build and test the basic receiver first; subsequently building and substituting the multichannel board, A2, for the basic oscillator on the A1 board.

Note that construction information in later pages gives data for covering any of four bands with the receiver: 6 or 2 meters or either low or high commercial band. Although a quality receiver, the cost of the parts involved allows the receiver to be very competitive with the cheapie monitors available on the market for public-safety monitoring, to say nothing about having a better one than your neighbor.

The receiver as described uses a one-watt IC audio amplifier, which should

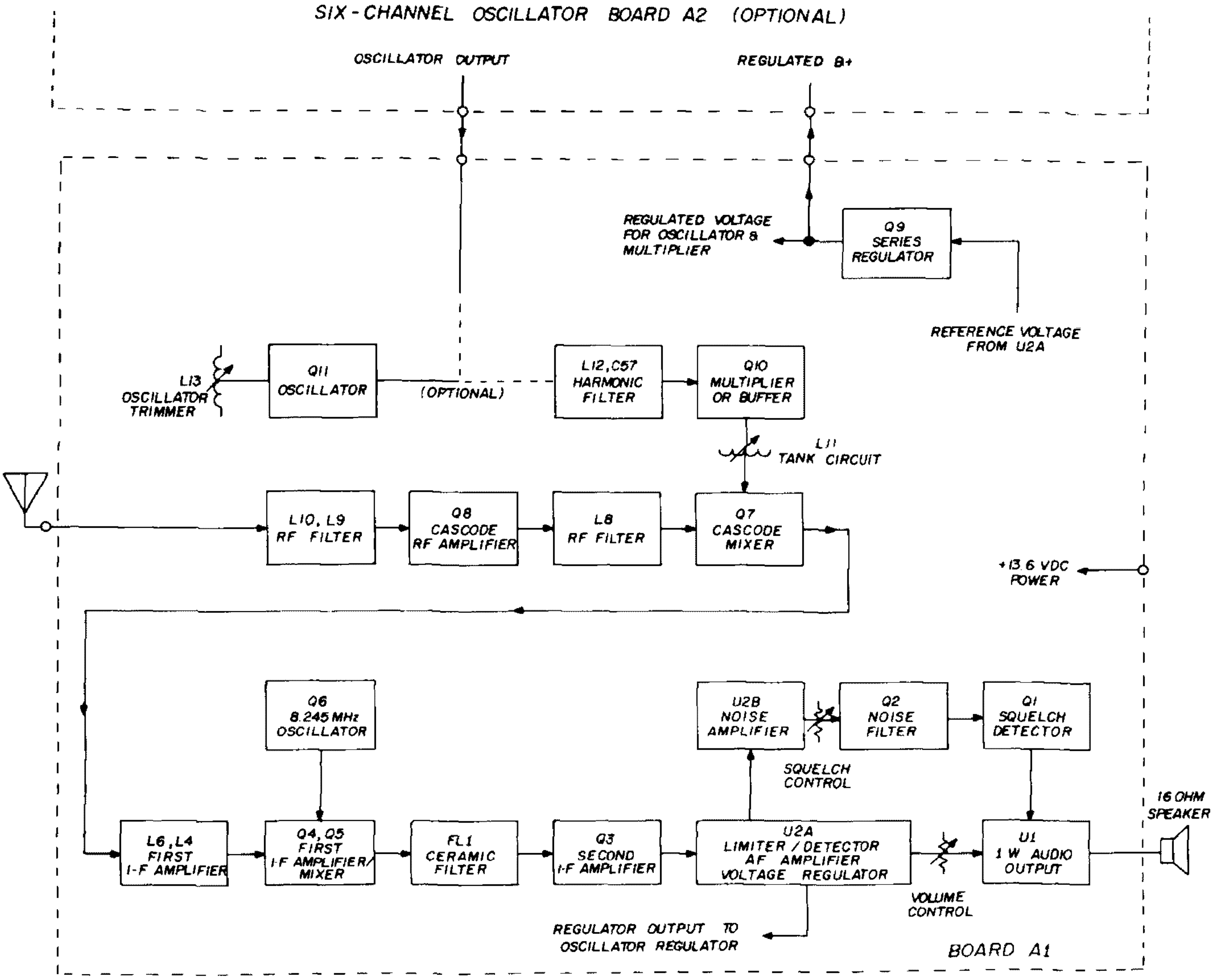


fig. 1. Functional block diagram of the vhf fm receiver.

suffice even for mobile operation. However, if desired, the receiver can be used to drive a higher power audio amplifier, such as the one I wrote up previously[2]. The same holds true for sensitivity. The receiver has been designed to be a good receiver for various applications, including requirements for minimum cross-mod in the front end. If you wish, a preamp can be added to obtain slightly better sensitivity. However, this is left optional, since you may be in a high signal level area where high front-end gain and selectivity against unwanted interference are mutually exclusive.

functional description

A block diagram of the set is shown in **fig. 1**. The rf signal from the antenna passes through two tuned circuits, L10 and L9 with associated capacitors, and is amplified by dual-gate mosfet Q8. The amplified rf signal is filtered again and applied to gate 1 of dual-gate mixer Q7. The injection to mixer Q7 is applied to gate 2, and rf signal is applied to gate 1, thereby providing slightly additional gain and better isolation. The output of Q7 is applied to first i-f filter L6/L4 to select the signal-minus-injection-frequency mixer product at a frequency of 8.7 MHz. This frequency was chosen to provide good operation at either high band or low band and to provide good image rejection with a fairly simple filter scheme.

Injection for mixer Q7 is obtained from Q10 at the optimum level of approximately 1-volt rms. Q10 functions as a tripler for high-band operation and as a buffer for low-band operation. The input to Q10 is applied either from internal oscillator Q11 or from multichannel board A2 through harmonic filter L12/C57, which is a series-tuned trap. This minimizes spurious mixer injections.

The oscillator crystal is trimmed with series inductance to allow a vernier adjustment of about ±1.5 kHz for netting to channel frequency. Using a coil rather than a variable capacitor provides a much easier method of adjustment because of the multi-turn range of a slug-tuned coil as opposed to a 180-degree capacitor swing. The B+ for the oscillator(s) and multiplier/buffer is obtained from series pass transistor Q9, which uses an existing regulated voltage inherent in the detector IC for a reference.

Note that part of the magic art of receiver building is the knowhow in bypassing each circuit element to common ground. Optimum values for each frequency are chosen for frequency of resonance to provide ideal rf grounding. Likewise, the decoupling between stages must be done very carefully to prevent random transfer of rf energy between stages on the B+ line. Such measures not only ensure that oscillation will be suppressed, but provide optimum gain in each stage. In some cases, multiple bypasses were used to provide grounding at more than one frequency. The selection of coupling and tuning capacitors in the front end is also a fine art, ensuring optimum transfer of signal without overcoupling or under-Q'ing tuned circuits.

The first i-f signal from L4 is applied to Q4/Q5, a cascode mixer/amplifier. Refer to the schematic (**fig. 5**) temporarily. Oscillator injection at 8.245 MHz is generated in dc-coupled oscillator Q6. This transistor is connected directly from the base of Q6 to the base of Q4. I don't know that this trick saves anything but two resistors and a capacitor, but I thought it was cute. (Might as well tell the truth!) Regency has a scheme using an MC-1550 coupled to a discrete oscillator this way. I thought it was a good trick, but I went a step further. A friend of mine convinced me that there is no excuse for using MC-1550s for a simple amplifier or mixer, so I replaced the one previously used here with two discrete transistors and a few resistors. You might want to try this trick in other circuits you build. Bypass the second base, and you have a neat cascode amplifier.

The second i-f output of 455 kHz from Q4/Q5 is filtered by a bandpass ceramic filter. This is one of the greatest little devices going! Essentially, it is an integrated-circuit Permakay filter, hardly any bigger than a dip IC. I chose one with nine poles (nine ceramic elements) which

provides a selectivity of ±7.5 kHz to allow operation on either narrow-band or quasi-narrow-band signals (as used by most hams nowdays).

The ceramic filter described is much smaller than a crystal filter and is considerably less expensive than the $30 approximate price of equivalent crystal filters. In addition, it is easy to match, requiring only a resistive 1500-ohm load at each end. Being composed of ceramic elements, no dc blocking capacitors are required either. The only drawback is that they must be imported so they are not readily available in single lots.

The filtered second i-f signal is amplified by Q3 and transformer coupled through an i-f can into U2A. Q3 is another attempt to stamp out needless ICs. An MC-1550, as originally used, adds nothing over a discrete transistor. The second i-f signal is applied to a 70-dB limiter amplifier in U2A, which consists of a series of differential amplifiers. The limited i-f signal is detected by a differential-peak detector in U2A. A separate stage in U2A provides a low-impedance audio-driver stage to feed the volume control and squelch circuit. Also incorporated in U2 is a series of voltage regulators which provide regulated +11 volts for operation of the IC. Some of the regulated voltage is tapped off to provide a reference voltage for the pass transistor in the front-end oscillator section.

Audio from the volume control is applied to an integrated circuit one-watt audio amplifier, U1. A complex series of de-emphasis networks is used in conjunction with U1 and Q2 to provide high-frequency roll off required for fm operation. The output of U1 is capacitively coupled to a 16-ohm loudspeaker provided externally. The 16-ohm load is optimum for a one-watt speaker driver amplifier such as U1, so you should strive to obtain a 16-ohm speaker and not use your old 3-ohm speakers, although they will work to some extent. They will draw more audio current than normal through U1, so the level should be held down if a low-impedance speaker is used.

An additional audio output from the preamp in U2A is provided to a separate section of U2, dubbed U2B, which is an audio amplifier normally used to drive a power transistor. Since this function was redundant to the requirements for operating U1, I used this section as a noise amplifier for the squelch circuit. Coupling circuits and a selective filter network in the emitter circuit of noise filter Q2 were designed to pass a narrow band of noise frequencies around 20 kHz. This noise component is present only in the absence of a carrier from an incoming signal, so it provides good reliable squelch operation. (This is not the case with the CA-3089, which operates on the presence of a directly detected carrier. That system allows any rise in signal level or noise level to open the squelch.)

The 20-kHz noise signal is detected in Q1 to provide a filtered dc signal to squelch trigger one input of audio amplifier U1 to provide quieting of the otherwise-present noise output when no signal is present. It is common practice to use a separate squelch gate to mute the audio path; however, my research has shown that it takes redundant components to do that with results which sometimes are not as good. Often, diode gates allow noise leakthrough when no signal is being received. Transistor gates operate fairly well, but in the same manner as squelching the IC directly.

detailed circuit descriptions

Now that your curiosity about the receiver is satisfied, let's take a look at the inner workings of the integrated circuits and their applications in the receiver.

Refer to fig. 2. The Motorola MFC-6070 one-watt audio amplifier is very similar to the discrete direct-coupled amplifier circuit developed by Motorola several years ago, using complimentary transistors. The primary differences are that all transistors are npn and that the input stage consists of a Darlington differential arrangement. Normally, the input signal is applied to pin 3 and negative

feedback is applied to pin 2. Pin 1 is grounded, and pin 5 has B+ applied. Pin 4 has B+ applied through a bootstrap arrangement to provide clean response at high input peaks. Two diodes in series between bases of the power transistors in both sides of the circuit set the bias

squelch control. When no signal is present, full B+ is applied to one side of the differential input stage on the MFC-6070 to effectively cut off the audio output by saturating the amplifier.

The MC-1358PQ, shown in **fig. 3**, is a country cousin of the MC-1357 and the

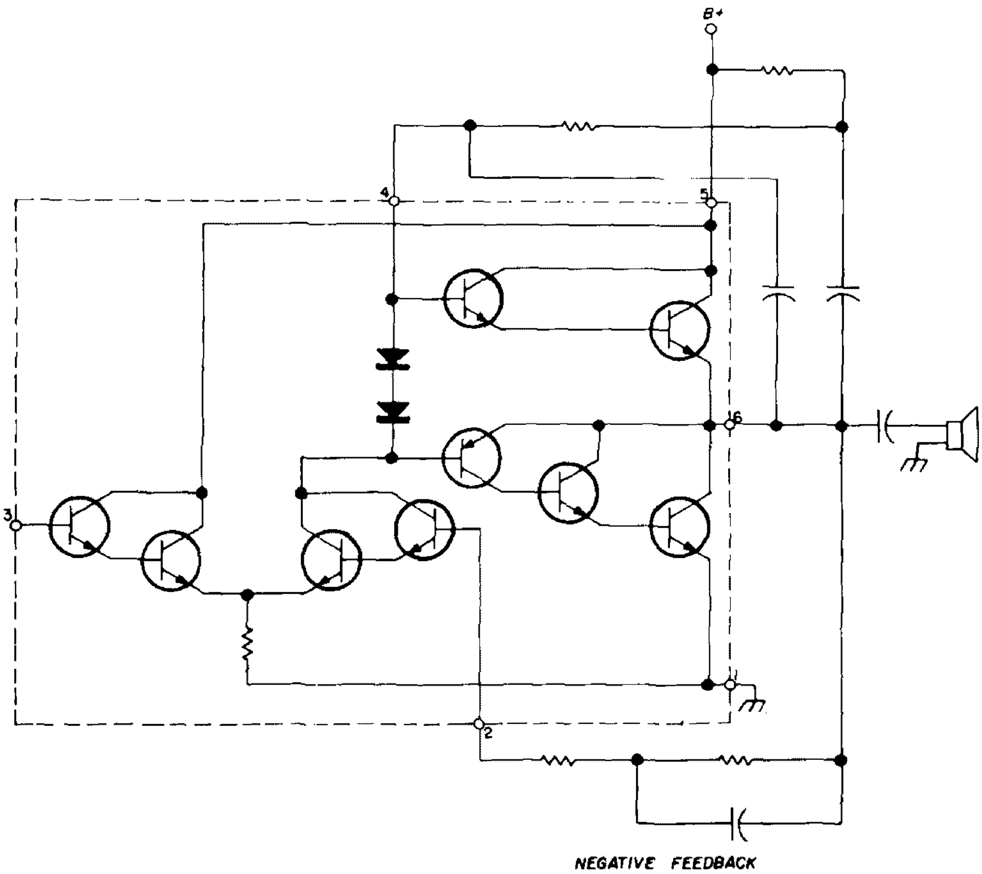


fig. 2. Basic schematic of the Motorola MFC-6070 one-watt audio amplifier.

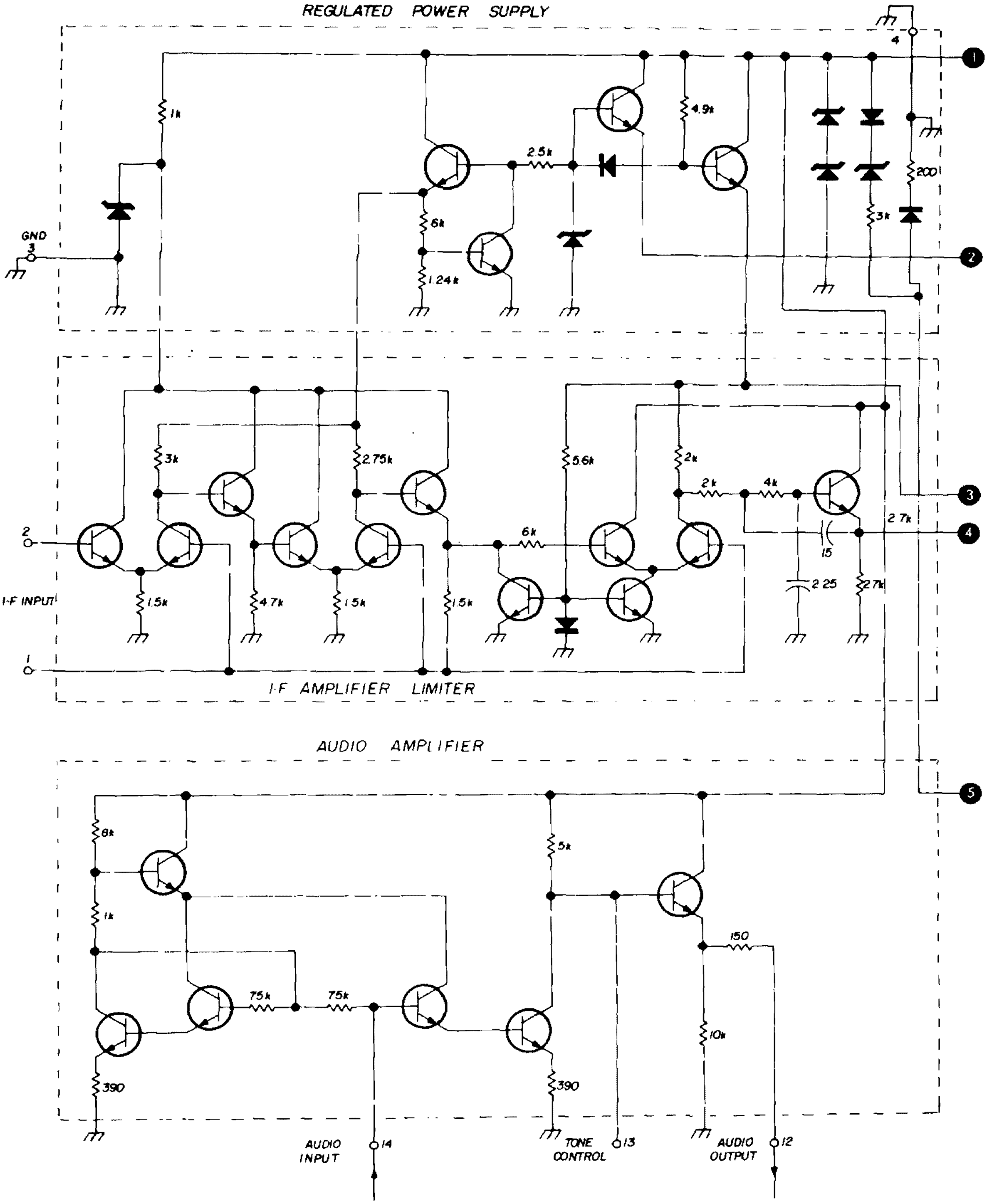
difference to prevent cross-over distortion.

A capacitor from pin 6 to pin 4 stabilizes the amplifier against high frequency oscillation. The output is capacitively coupled (no transformer required) to a 16-ohm speaker load. In this case, the speaker was returned to ground for convenience, although it is common practice in consumer equipment to return the speaker circuit to B+.

Biasing is applied to pins 2 and 3 in approximately equal amounts for normal operation. In this application, I returned the biasing circuit for pin 3 (not shown) to the squelch detector to provide

ULN-2111A integrated circuits which have been around for a few years. Although the former and latter are priced the same, the MC-1358 has all kinds of added *goodies*. One added feature is an array of voltage regulators as shown in the upper left corner. B+ is applied to pin 5 through R_S, which drops the +13.6Vdc to the +11Vdc established by the shunt regulator consisting of a pair of zener diodes. Other devices in the section supply various reduced bias voltages to the individual signal stages.

The input i-f signal is transformer fed to pins 1 and 2, which provide a differential input port. The signal is amplified



and limited by a four-stage 70-dB amplifier arrangement and applied to the detector section. Unlike the quadrature detector used in earlier ICs, the MC-1358 uses a differential peak detector. The signal is peak detected in two sides of a differential circuit. The 15-pF capacitors incoroporated in the circuit are actually part of the peak detection process.

The signal is applied directly to one side of the detector and is applied to the other side through a tuned-circuit ar-

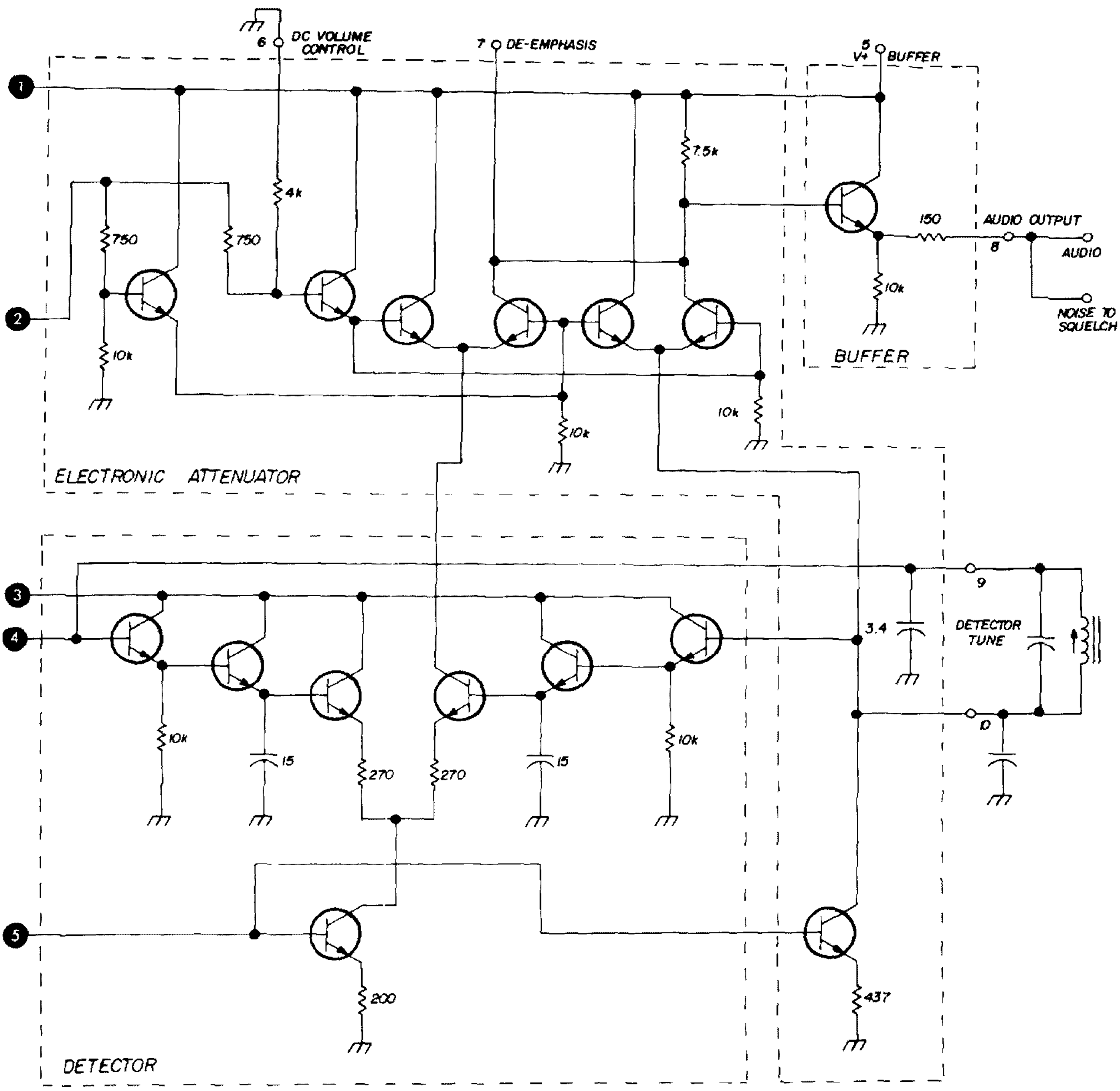


fig. 3. Basic diagram of the Motorola MC-1358PQ sound i-f circuit.

rangement to provide a frequency-sensitive component to shift the level (not the phase) of that signal. Thus, depending on instantaneous frequency, the inputs to the two sides of the detector will be different and produce an output difference. As the frequency changes, the relative difference of the two inputs changes, causing an audio output signal to be generated. The tuned circuit consists of a pi-network, employing an internal 3.4-pF capacitor on one end and an external 20-pF capacitor on the other end to provide a slight impedance transformation from pin 9 to pin 10.

The audio output from the detector is applied to an electronic attenuator section. In television usage, this section normally provides about 80 dB of audio attenuation variable by connection of a 50k dc potentiometer to pin 6 of the IC. In this application, though, I couldn't take advantage of this feature, since reducing the signal level at this point would also reduce the noise output to the squelch circuit. This section and the

adjacent buffer do provide the desired audio output level and low-impedance output, however, and partial de-emphasis is performed at this area by connection of a capacitor from pin 7 to ground. Full de-emphasis is not done at this point, though, since you need considerable high-frequency noise output to operate the squelch circuit.

audio from driving a high-gain noise amplifier, since this would result in clipping-produced distortion which would "pump" the squelch circuit during normal voice reception.

channel crystals

Channel crystals, Y2 for A1, or Y1 through Y6 on multichannel board A2,

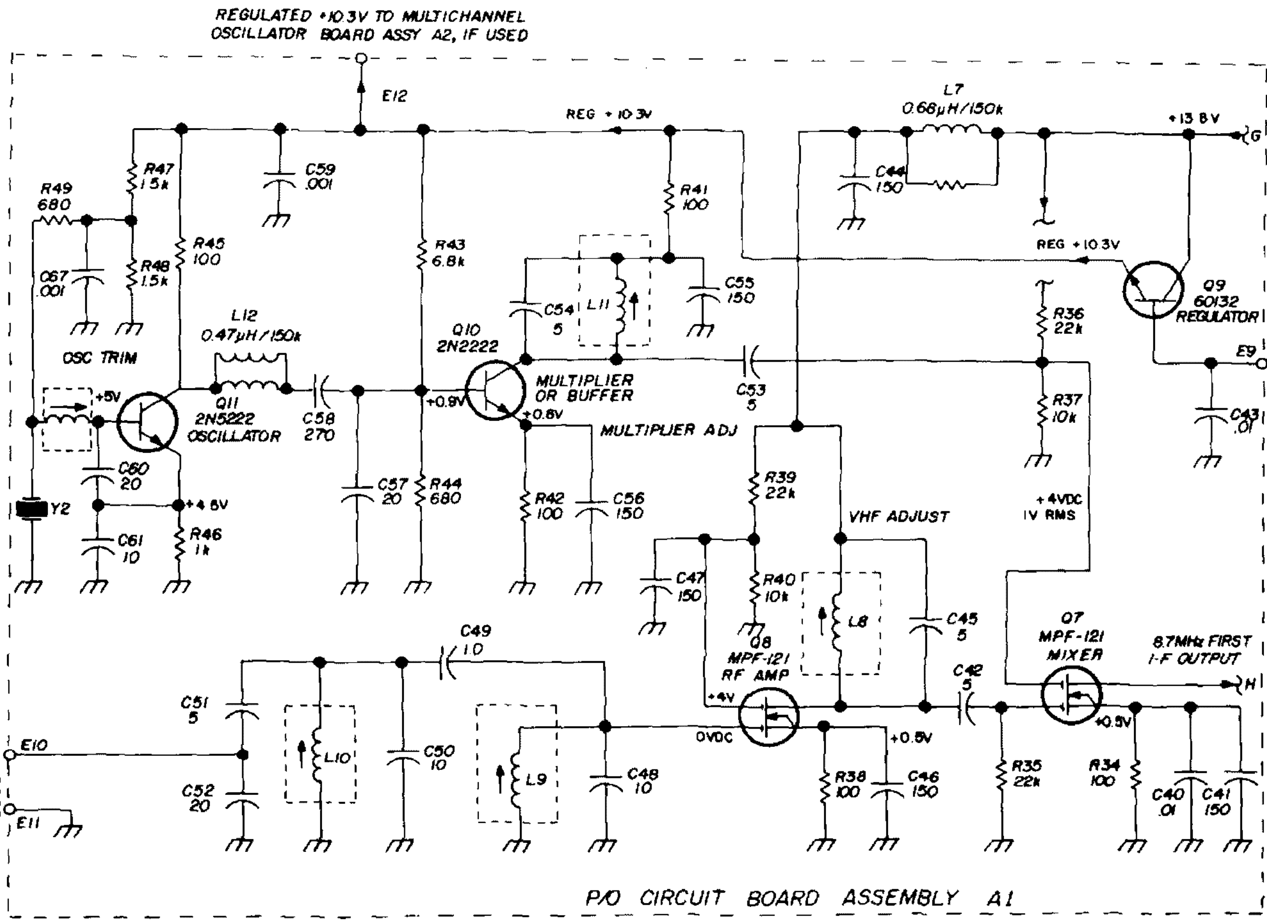


fig. 4. Front-end section of the vhf fm receiver. Coil and capacitor data is given in table 1.

The audio amplifier section at the lower left corner is intended to be used for driving the low-impedance input at the base of a power transistor. However, not needing such a low impedance and high drive level for the MFC-6070, I used this amplifier section as a noise amplifier for the squelch circuit. The three stages provide about 40 dB of gain, and small coupling capacitors were used to begin the low-frequency rejection process in feeding the noise-filter stage. Care must be taken in designing noise amplifiers to prevent high levels of low-frequency

are third-overtone, series-resonant types in HC-18/U (miniature, wire lead) holders. The formula for operation on 2 meters or the adjacent high commercial band is:

$$F_{xtal} = \left(\frac{\text{channel frequency} - 8.7\text{ MHz}}{3}\right) - 1\text{kHz}$$

For 6 meters or the adjacent low commercial band, the formula is:

$$F_{xtal} = (\text{channel frequency} - 8.7\text{ MHz}) - 1\text{kHz}.$$

One kHz is subtracted from the final calculated frequency to allow the pulling

range of the crystal to be centered at the channel frequency, since the system tends to have much more range at the high side than at the low side of the true series-resonant frequency.

Various types of crystals are available, and almost any good crystal house can supply them. The type desired depends on the use of the receiver and the environmental temperature range which will be encountered. For monitor receiver use in the shack, I would recommend monitor set crystals, available for about $5.00 from many sources. My firm, Hamtronics, sells Crystek's monitor crystals, since they are available quickly on a crystal certificate basis and are mailed directly to the customer. For operation over a wide range of temperatures, as in mobile operation, I would recommend the standard commercial type of crystal, which has a specified temperature curve. They cost around $6.50. The circuit itself is stable, so the overall stability with temperature variations will depend almost entirely on crystal quality. If you will be in the shack all of the time, though, I wouldn't bother spending extra money for the more expensive crystals.

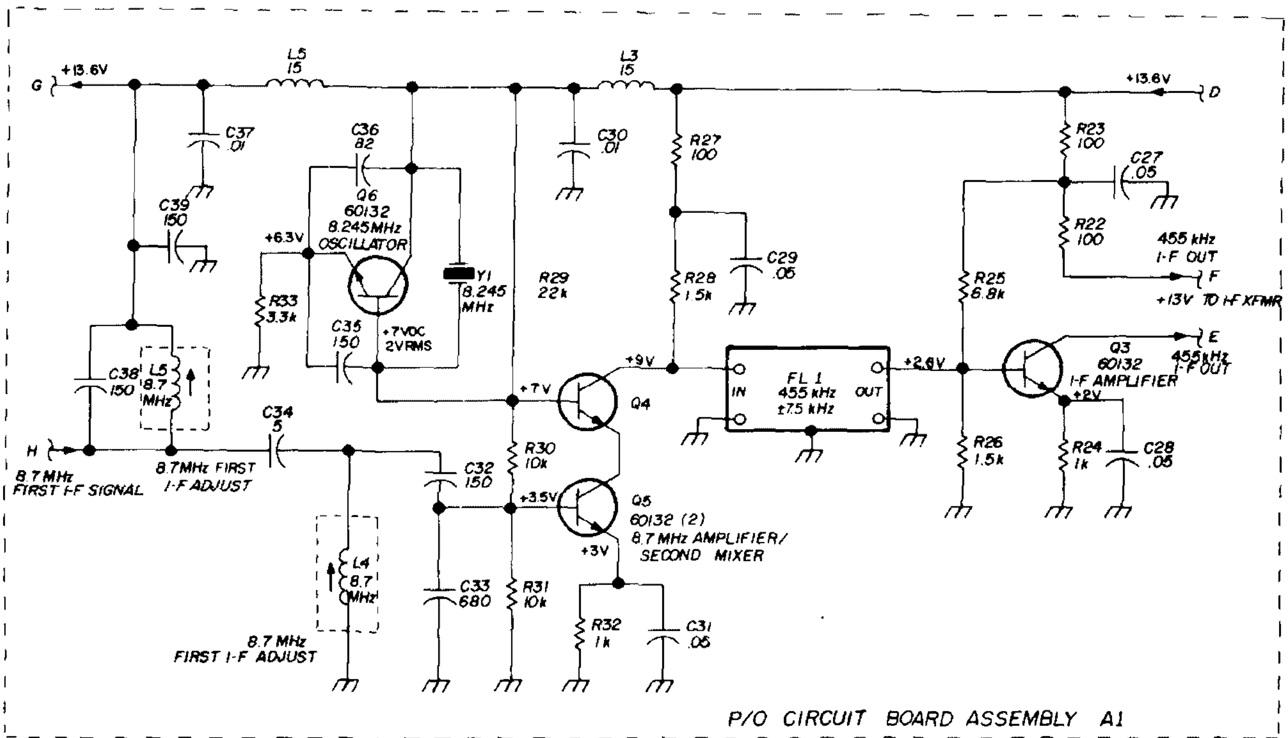


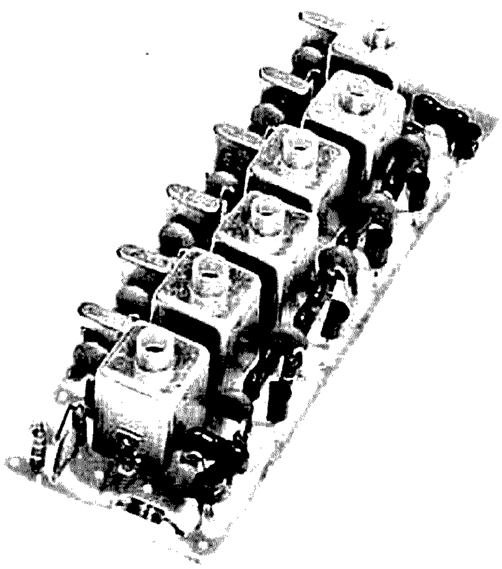
fig. 5. I-f section of the vhf fm receiver, including the first and second i-f filters. Filter FL1 is a Japanese ceramic filter with a bandwidth of 14 kHz.

semiconductors

At this time, there are no direct substitutes for the integrated circuits used in this receiver, although there are a variety of different audio output ICs available which could be used with different connections. In discrete devices, the 60132 is our house-numbered general purpose npn high frequency silicon transistor; it can be replaced by any good 200-MHz F_T unit. The 2N5172 is a very attractively priced transistor, having a single lot price tag of about 16 cents. The 2N5222 is a 450-MHz silicon type which is specified with closer tolerances than the 2N5172, although it is a similar type with a fairly low price. The 2N3644 is a high-gain 60-MHz unit of the pnp silicon type. In the applications in this receiver, any pnp silicon type with high gain at i-f frequencies can be used.

Unless otherwise specified in the schematic diagrams or in the text, all resistors are ±10%, ¼ watt; all capaci-

tances marked in whole numbers are in picofarads and those specified in fractional numbers are in microfarads; all coil inductances are in microhenries. Dc test voltages for various idling conditions are



The six-frequency crystal deck.

marked on the schematics at significant test points for troubleshooting purposes.

coils and chokes

Before assembling the receiver circuit board(s), it is necessary to wind the rf coils, high i-f coils, and one or two rf chokes. Data for coils is given in table 1, and the coil winding technique is illustrated in fig. 9. Winding your own coils has two advantages. First and foremost, is the cost savings. Second is the fact that you can do as well as the finest inductors commercially available.

Rather than buy an expensive off-the-shelf coil form, we have ours custom made by the largest houses in the industry. We use the types basically made for commercial two-way sets. I say houses because the form is made by one manufacturer, the slugs by a second, and the shield cans by a third. Unfortunately, these people don't even like to deal with us for 10,000 pieces at a time, but we manage to have them made to our specs regardless. We do this so you don't have to struggle with fragile ceramic forms with loosely fitting slugs. The forms used, by the way, are 10-32 types of 0.8 inch length. Cans are half-inch square by three-quarters high with Berg solder lugs. The slugs are 10-32 x 3/8 inch long with a small hex slot.

Coils are wound as shown in fig. 9. Soldereze number 26 wire is used so that insulation is thermally stripped with your soldering iron. Carbonyl – J slugs are used for all applications except the 8.7 MHz first i-f coils, L4 and L6, which use slugs of carbonyl TH material identified by a different color. To wind a coil, insert about a half inch of wire through any of the six plastic funnels in the base of the form. Bend the wire at right angles to the base, and begin winding with turns tightly spaced. After the proper number of turns is applied, make a right angle bend in the wire, trim the end to allow a half inch protrusion, and insert the wire through the funnel opposite the starting funnel. All coils are specified in half turns, so that windings finish 180 degrees around from the starting funnels.

After the coil is checked for proper number of turns, perform the following magic. Lock a tuning tool in your bench vise, and insert the slug in the form. Slip the coil and slug over the tuning tool so that the coil is stationary with the funnel end of the base outward. Using a hot soldering iron, tin the insulated Soldereze coating on the wire, starting at the tip of the wire where there is no insulation at the end. As the wire heats up, the insulation will melt off, and the wire will tin with the solder applied to the iron.

Continue up the wire to within about 1/32 of an inch of the form. You will have to remove the iron quickly when solder reaches that point. When done, the plastic at the funnel will melt slightly and bond the wire to the form, making a neat coil assembly. The other lead should be done in the same way. The coils are eventually mounted at the bottom by the leads being soldered to the circuit board, and the coil tops are held rigid by the tops of the shield cans, which slide over

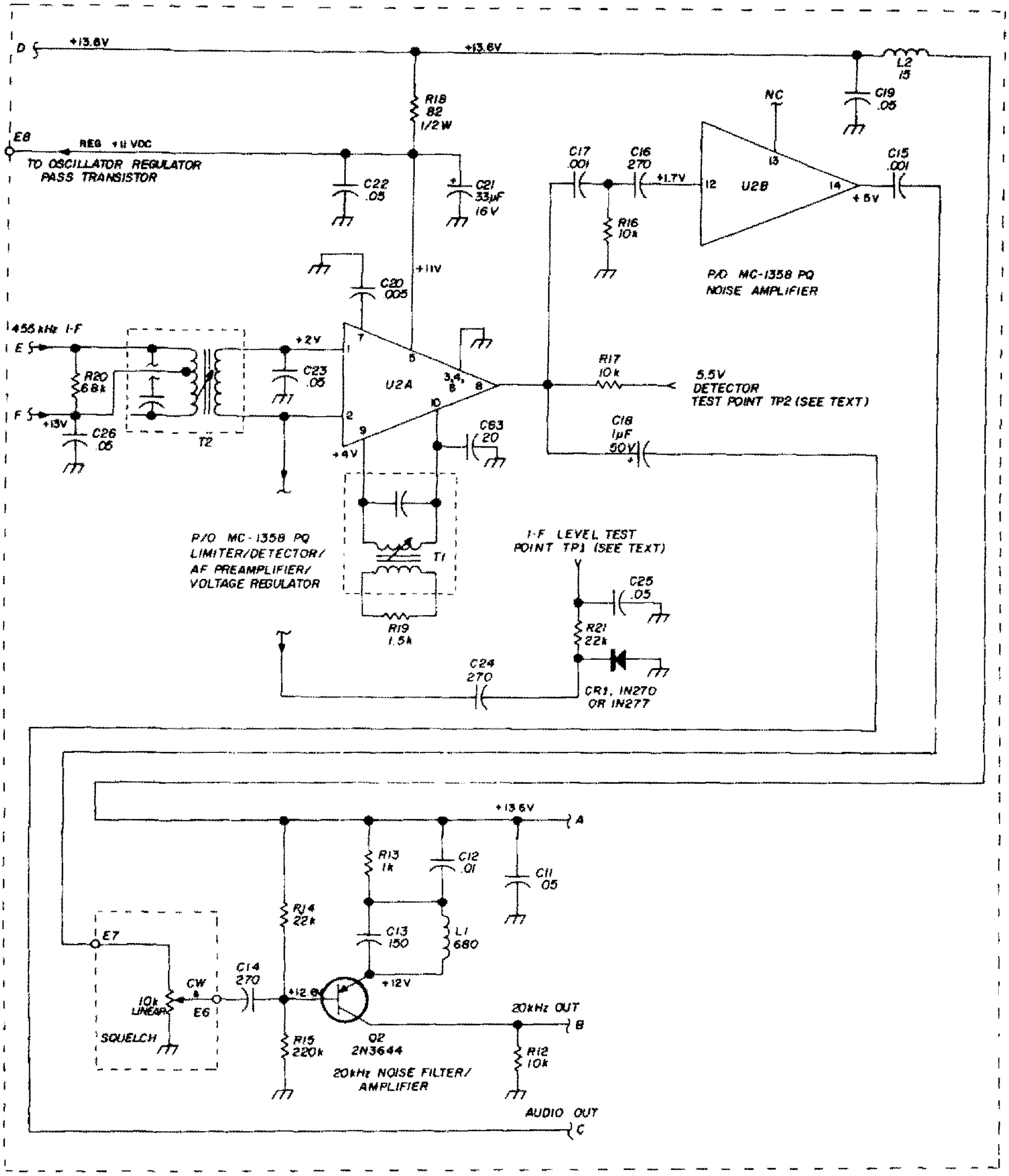


fig. 6. Limiter/detector section of the receiver. Note the test points provided for alignment. The squelch control is external to the circuit board.

the tops of the forms. This gives you a professional, modern coil assembly which will amaze you. Our preamp customers have been writing for parts so they can use these coil forms and shields in other projects.

I have started using hand-wound 0.47- and 0.68-μH rf chokes, since paying a lot of money for a part so easily made goes against the ham instinct, once you know the tricks. Fig. 10 shows you how to make these. Chokes are wound with number 28 Soldereze wire, which is the smaller size supplied with the kit. A 150k

resistor was used, although any ½-watt resistor with a high resistance will work well as a form. Wrap two or three turns around the lead near the body at one end. Wind on the 17 or 20 turns, which will fill most of the length of the resistor body, and then terminate the other end in the same manner. Twist the wire around the body tightly, and then solder away the insulation to make the connection to the resistor leads. You now have an rf choke!

For those of you interested in making other values for general use in your own projects, you can scale the turns as being in proportion to the square root of the inductances. The tricky part is that the series self-resonance must be well below the operating frequency. I use a Q-meter to check inductance and a vhf signal generator (HP-608) and an rf voltmeter to check resonance, but a good grid dipper can be used (if you can find one good over 100 MHz).

board layout

If you decide to lay out your own pc board(s), general vhf layout techniques should be used throughout. Even the integrated circuits, although operating at 455 kHz and audio, require special layout precautions due to the high impedances and high open-loop gains involved. Single-sided boards work well if layout is good and sufficient areas of ground plane are provided; however, stay away from board materials with poor dielectric properties. General layout can be seen in the photograph.

Since doing a good tape-up can take several weeks of effort, I suggest that you write if you only need one or two boards, and we can probably help you out. It is most practical to obtain the kit, however, to get the coil forms and filter as well as the board(s).

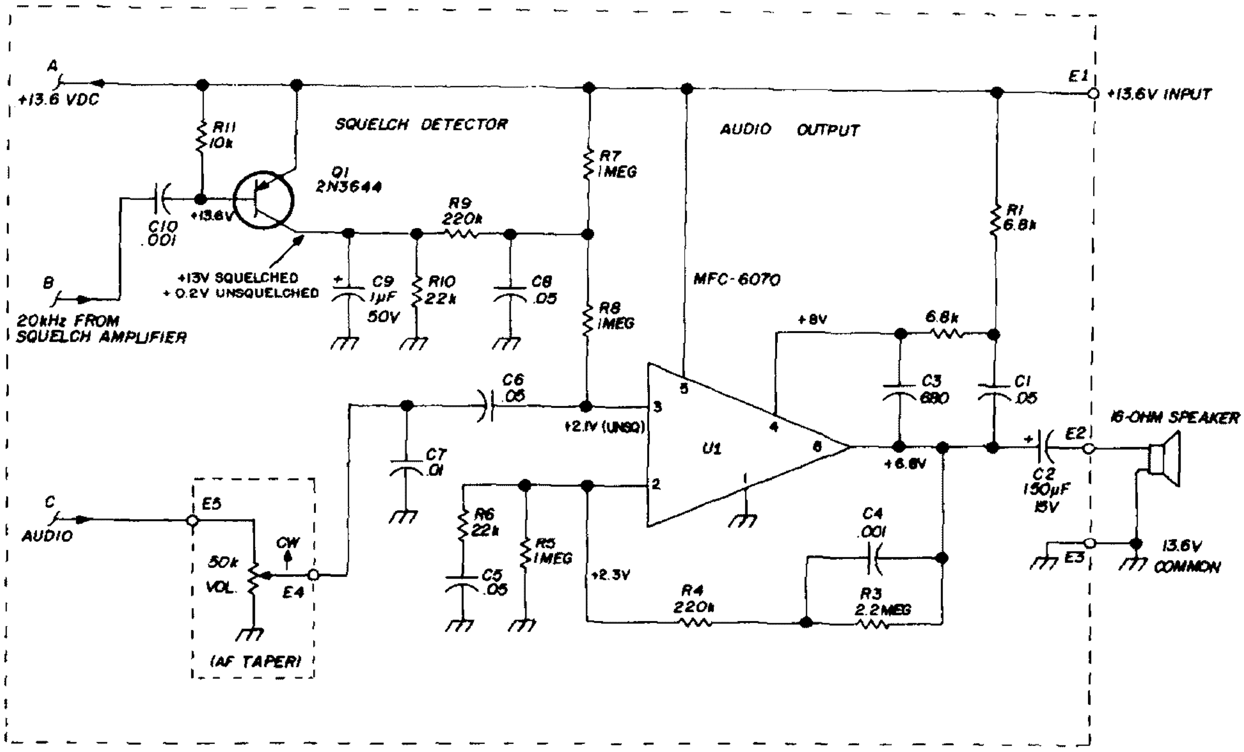


fig. 7. Squelch and audio section of the receiver. Volume control and speaker are external.

construction sequence

I have a philosophy about building a project like this which has been developed by hard knocks. *In general, when* building anything, I progress one or two stages at a time, testing as I complete each section. Things do go wrong: a

wrong part in a certain place, a bad solder joint, or even a defective part. It is much easier if you build and test in sections, verifying the proper operation up to a certain point before introducing more variables. Anyone who has built some project only to find that it oscillates (assuming it shouldn't) can tell you how hard it is to narrow it down if this plan isn't followed. Also, with receivers, I find it is always best to start at the speaker and work backward. You may have noticed that even the component reference designators are numbered in that direction.

With the section-at-a-time approach in mind, the schematics have been carefully broken down into convenient functional sections and pc board A1 had been laid out to allow testing after building each section as shown. You should start with the audio section in fig. 7 and work toward the antenna. In that way, the speaker and audio output stage will act as an extra test aid so that you can hear what's happening as you progress. For testing each section, a signal generator can be connected to the components at the left side of the schematic (sometimes through a blocking capacitor).

As a matter of fact, by the time you get the high i-f working, you can connect an antenna and possibly hear some short *wave stuff. (You remember what that is!)* When you get to the front-end section in fig. 4, you should mentally break the schematic into two smaller sections, and build and test the oscillator and multiplier (or buffer) first. If you are planning to use an A2 multichannel board, you should either build and test that first, or better, build the single-channel arrangement on the A1 board for initial testing purposes and make A2 a separate project. *Life is simpler that way!*

construction notes

Following are some practical suggestions to consider when assembling your receiver.

1. Check and double check the direction in which the various components should be installed. In particular, take care when inserting the leads of transistors, ICs, electrolytic capacitors and the ceramic filter. The filter which I used is said to operate ok when reversed but with somewhat more insertion loss. If you use the Hamtronics kit, note that resistors mounted in a stand-up arrangement may be oriented with a lead coming off the top facing a certain pad. (Drawings with the kit show preferences.) A preference may be to protect from shorting to adjacent leads, or in the case of TP1 and

table 1. Coil and capacitor data for the vhf fm receiver.

	2 meters	high-band	6 meters	low-band
L4	28½ turns	28½ turns	28½ turns	28½ turns
L6	26½ turns	26½ turns	26½ turns	26½ turns
L7	0.68 μH	0.68 μH	5.6 μH	5.6 μH
L8	3½ turns	3½ turns	9½ turns	9½ turns
L9, L10	2½ turns	2½ turns	8½ turns	8½ turns
L11	3½ turns	3½ turns	10½ turns	11½ turns
L13	11½ turns	11½ turns	11½ turns	12½ turns
C39, C41, C44, C46, C47	150 pF	150 pF	680 pF	680 pF
C42,C53,C54	5 pF	5 pF	15 pF	15 pF
C45	5 pF	5 pF	20 pF	33 pF
C48	10 pF	10 pF	20 pF	33 pF
C49	1 pF	1 pF	2.2 pF	2.2 pF
C50	10 pF	5 pF	10 pF	15 pF
C51	5 pF	5 pF	20 pF	20 pF
C52	20 pF	20 pF	50 pF	50 pF
C55, C56	150 pF	150 pF	0.001 μF	0.001 μF

TP2, the top resistor lead provides a convenient test point connection.

2. In this day and age, you needn't worry too much about heat sinking semiconductors. Just make sure you use a good, clean pencil-type soldering iron, and make your solder connections quickly. learned the hard way by destroying the coil forms in about 50 preamps because I wanted to wash them nicely with trichlor.) If you use good solder, you shouldn't have to worry about cleaning the board except for appearance sake. (I find that popular multi-core solder leaves

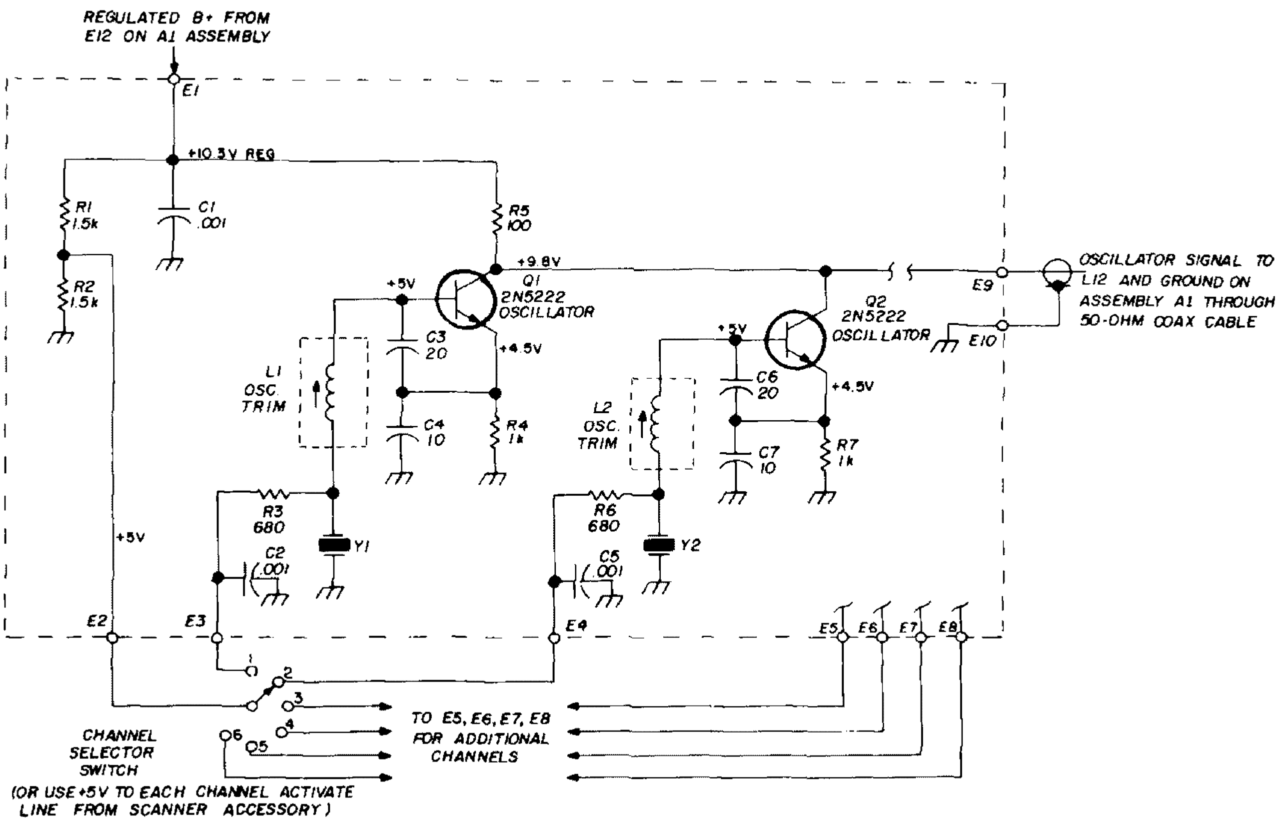


fig. 8. Circuit board A2 which may be used to provide multichannel operation of the vhf fm receiver. Six channels are included on the board, but only two are shown here.

Generally, pc board land areas can become separated from the board material long before a semiconductor device is damaged. Nowdays, most parts are made to allow flow soldering in mass production, so think accordingly. If you want to worry about your semiconductors, check the transients on your power supply and your polarity when making connections. These are the things which hams forget to check until it is too late!

3. If you wash your finished board off with solvent to remove excess flux, watch out! Plastic parts can become swollen and deformed with solvents such as trichloroethylene. Freon is probably safest. (I the darkest, messiest flux residue, so I threw mine out.)

4. After coil forms are soldered securely in place, the shield cans should be installed over them; and the lugs should be soldered to the board, both to make good ground connections and to provide mechanical rigidity.

testing and alignment

As previously suggested, it is advisable to build and test a section at a time, beginning with the audio portion in schematic fig. 7 and working backwards toward the antenna. The following test equipment should be available.

1. Vtvm or fetvm, preferably one which has a dc range that can indicate full scale values of about 250 mV.

2. Vhf and hf signal generators left on long enough to be fairly stable. No modulation required. An hf signal generator will suffice if harmonic frequencies can be determined fairly well.

3. Power supply, regulated or unregulated to provide +13.6 Vdc filtered fairly well with a current rating of 500 mA. Be sure the power supply has no voltage transients.

4. Rf probe for vtvm. See ARRL handbook for an easy one to build if you don't already have one. Works fine! Required for checking oscillators.

5. Frequency counter for vhf range — optional to check oscillator frequencies. Can also be done with on-the-air signal checks.

To augment alignment procedures, two test points have been designed into the receiver, including an rf level detector for measuring relative signal levels at the i-f stage. This is roughly equivalent to checking limiter current in a tube-type receiver. This is a novel feature not often found even in professional commercial radios.

A completed receiver normally draws about 75 mA (1 watt) when squelched and about the same amount unsquelched with the volume turned down. Current increases rapidly as audio level is increased, but shouldn't exceed 300-400 mA on peaks. Of course, when testing the first section by itself, current drain will be relatively low.

Begin testing the audio circuit in fig. 7 by connecting the external speaker, volume control and power to the board as shown. Check the current drawn from the power supply to be sure that it is less than about 50 mA. At this point in construction, the squelch circuit has no drive, so the audio output stage should be operative. You should be able to hear hum if you touch your finger to the volume control at E4 or E5, and you should be able to hear an audio signal (from any radio or audio generator) patched into point C on the schematic (top of volume control).

No appreciable distortion should be evident unless you have an ac ground loop from the audio source causing superimposed ac garbage. You should be able to measure the dc voltages marked at terminals of semiconductors on the schematic. Values aren't critical and vary with component tolerances and line voltage; so don't get shook if your values vary somewhat from the ones marked on the schematics.

With respect to the MFC-6070 audio amplifier, note that Motorola states that because of the high open-loop gain, it may be necessary to install an oscillation suppressor for high frequencies if the speaker impedance or very long leads to the speaker cause a reactive load of certain types to be applied to U1. I have not encountered this problem. However, I will mention the fix just in case someone should have need for it. A series R-C network, consisting of a 10-ohm resistor and a 0.1-μF capacitor, is connected from output pin 6 of the IC to pc board common to correct an oscillation.

After completing the limiter/detector section shown in fig. 6, you can connect a 455-kHz signal generator through a 0.05-μF blocking capacitor to point E on the schematic with the shield connected to pc board common ground. The nominal level for testing should be about 5 mV at first. Temporarily connect the dc probe of a vtvm to pin 5 of U2. Be careful not to short to adjacent pins. Measure and record the regulated dc voltage at this point. Nominal level of +11 Vdc can vary according to the tolerance of zener diodes in U2. Divide the measured voltage by 2. The calculated voltage (about +5.5 Vdc) is roughly equivalent to zero on a discriminator meter when measuring the voltage at TP2, the top of resistor R17. The direct-coupled emitter-follower audio-output stage in U2 will present a variable dc voltage dependent on i-f signal frequency after alignment. With zero-deviation (center frequency) input signal at 455.0 kHz, the voltage is about half the regu-

lated supply voltage.

With the 455.0-kHz signal applied, connect the vtvm probe to TP2, and adjust T1 through its range. You should observe a dc voltage variation of about ±1.5Vdc as you tune through the range. Now set T1 for a value equal to the calculated test voltage previously determined. Then, scan the signal-generator frequency through about a 30-kHz range around 455 kHz, and note the S-curve effect and the low and high voltage levels reached. Find and record the mean voltage (center of the spread), which should be close to that previously calculated. This is your new "on-frequency" reference voltage to be used for alignment in future procedures. Set the signal generator to 455.0 kHz, and adjust T1 to obtain the new test voltage at TP2. This completes detector alignment for now.

Note at this point that squelch operation and alignment of T1 cannot be checked due to high-frequency stages not being completed. The squelch can be checked later, when sufficient noise level is available from front-end on through high and low i-f stages. At that time, the squelch action should be apparent as the squelch control is rotated. Normal operation provides complete noise silencing with the control clockwise approximately 25-50%. Squelch action should be abrupt as in any good fm receiver.

After completing the portion of the receiver in the i-f section (fig. 5), connect the dc probe of your vtvm to i-f test point TP1 (top of R21). Set the vtvm to lowest possible positive dc voltage range, preferably in the millivolt range if you have one of the newer fetvms. The meter will indicate relative i-f signal level once the threshold of CR1 conduction is reached, which is about the same you would expect from an rf probe connected in the i-f circuit. Apply a 455.0-kHz signal through a 0.05-μF blocking capacitor to the base of Q3. Adjust the signal generator level to obtain a meter reading of about ¼ scale. Then, align T2 for maximum meter indication.

Connect the vtvm rf probe to R29 or R30 at the base of Q4, and check the Q6 oscillator level, which should be about 2 Vrms. Allow for inaccuracies in your rf probe. The important thing is to note that oscillation begins immediately as the power supply is turned on.

Reconnect the signal generator through a 0.01-μF blocking capacitor to point H on the schematic. Set it for about

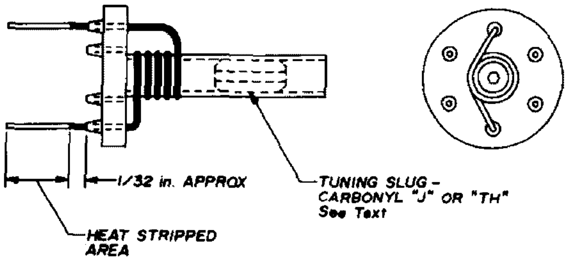


fig. 9. Rf coil winding and stripping. Data is given in table 1. Although turns are shown spaced out, they should be wound tightly together.

1 mV output (increase later if necessary) at 8.7 MHz. Connect the dc vtvm probe to TP1, and set the meter for lowest positive dc voltage range. Adjust the signal generator to obtain partial scale deflection, and adjust frequency for a peak. Then, adjust first i-f coils L4 and L6 alternately for maximum. Keep the generator level as low as possible while still maintaining sufficient meter deflection required for tuning. (Be sure to use the correct tool, such as the small tip on most Heathkit tools, to prevent cracking the tuning slugs.)

Note that these coil forms, with turns wound starting right at the base, will not allow you to obtain two peaks, which is the normal test to be certain that you are peaking and not just running through maximum inductance. You will get only one peak on these coils. However, if you don't get a definite peak but just run out of range as the meter is rising, you must adjust the number of coil turns to allow peaking to occur. This is not likely to be necessary with L4 or L6; however, it may be necessary with front-end coils, especially if you want to operate at a frequency far removed from the design frequencies.

You should add one turn (rewind coil, using a different pair of funnels on form)

if the slug wants to be all the way at the bottom of the form, and you should take one turn off if the slug wants to be all the way out of the form. The latter is best done by heating the coil connection, under the pc board, corresponding to the highest lead on the coil form and pulling the lead out of the top of the funnel while the plastic around the lead is warm. Then, remove one turn, clip off excess wire, and reinsert insulated wire through funnel and board. The insulated lead can be heat-stripped right in place at the board's pad.

After completing the high i-f and low i-f stages, it is recommended that you build the regulator, oscillator and multiplier/buffer in **fig. 4.** Note that a jumper must be installed from pad E8 in **fig. 6** to E9 in **fig. 4** to carry the regulator base input signal. A jumper is used to eliminate the need for long bus wires printed on the board to prevent stray signal coupling along the printed wiring, and to allow maximum use of ground planes on the board. Set the oscillator coil to midrange, and connect the rf proble of your vtvm to C53 at R36 and R37.

Apply power, and adjust L11 for maximum, which should be at least 750 mVrms. If you have a frequency counter, you can observe the effect of L13 on oscillator frequency by the connecting counter at the base of Q10. Otherwise, *you should wait until the front-end* is completed and set the oscillator frequency with an incoming signal. Just make sure that L13 is set at a position in which the oscillator starts; a setting which should not be critical at all, since L13 has been designed to allow oscillation over the entire range with crystals of good activity.

If you have constructed the multichannel oscillator board A2, shown in **fig. 8**, it should be connected via miniature 50-ohm coax to the A1 board and tested when you decide to tie it in. Connect the center conductor and shield to A2 at pads E9 and E10 respectively. At A1, the Q11 oscillator circuit components shown at the left of the breaks in the signal and B+ lines in **fig. 4** should be removed from the board, if installed earlier, and the coax cable should be connected to the pads normally used for the collector of Q11 and one of the closer common ground pads. A number 22 hookup wire should be connected from E12 on assembly A1 to E1 on A2 to carry the regulated B+.

The desired channel control line (E3-E8) on A2 should be activated by

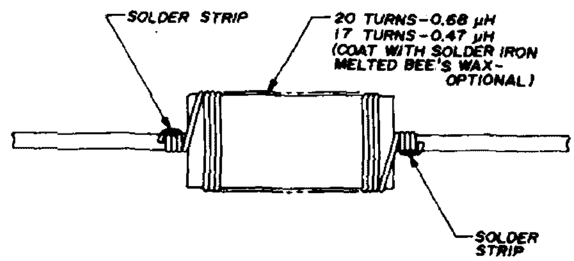


fig. 10. Rf chokes used in the receiver are wound around a ½-watt resistor.

either direct- or switch-connection to +5 Vdc terminal E2. The oscillator can then be tested as previously stated for A1-Q11. Note that the multichannel board can be used with any scanner project you wish to build which is capable of supplying about +5 Vdc to turn on the oscillator stages. Depending on the scanner circuit, you can get squelch voltage from any of a few different places on the receiver around the collector of Q1 in **fig. 7.**

The final step in construction is to assemble components for the front-end rf amplifier and mixer stages. Be sure that you install the coils in the proper positions according to table **1.** Note, too, that certain capacitor values change when using 6 meters or the low commercial band. Values shown on the schematic are for 2 meters or high band, and alternate values are shown in **table 1.**

To test the front-end, assuming that the rest of the receiver is operative, inject a signal-generator signal at the proper frequency at E10 and E11 on A1, and tune the coils for maximum with the dc probe of a vtvm *connected to* TP1 (**fig. 6**). Keep the signal level as low as possible while maintaining sufficient meter indication to tune by. As stated earlier, depending on exact operating frequency with respect to design center, it may be neces-

sary to change one or more coil windings by one turn to provide proper peaking within range of the tuning slug.

The antenna lead for the final packaged board should be connected from E10 and E11 with miniature 50-ohm coax to a connector you supply on the chassis, stripping the coax ends as short as possible. Long unshielded leads have series inductive reactance which will reduce the sensitivity of the receiver by dropping signal strength enroute to the input coil. A good method of connection is to wrap a length of number 22 bare wire around the shield and insert the end into the pad on the board. This method is neat and allows shorter center conductor connections.

The final alignment procedure consists of netting the oscillator(s) to channel frequency by monitoring at TP2 (fig. 6) for predetermined test voltage. Adjust oscillator coil(s) for that dc voltage while receiving or injecting a signal on-frequency. You may also do the adjustment for minimum distortion of the audio signal if desired, although it's not as accurate. You should then connect the vtvm to TP1, and adjust L11, L10, L9 and L8 in the front end and L6 and L4 in the high i-f for maximum.

T2 and T1 should be adjusted only with a signal of exactly 455.0 kHz injected into the low i-f, since both coils affect linearity of the detector response. Since these have already been adjusted, it is recommended that they not be touched. They should definitely not be used to net the receiver in lieu of the oscillator coil, since T2 and T1 must be aligned to the same center frequency as the ceramic filter to prevent distortion.

When completed, the receiver should provide 20 dB of audio noise quieting with a carrier signal level of about 0.3 μV applied, and the squelch should open with less than 0.2 μV. It is difficult to establish these levels though, as even good commercial signal generators have some leakage at these levels. It is especially difficult when the board *is unshielded*, although these is no real necessity to shield the board for normal operation. Boards may be mounted with small standoffs to a chassis or to the rear of a rack panel, if desired, to finish the installation.

mobile noise

A final note on mobile operation is appropriate. Ignition and alternator noise are a big headache to all concerned, but they needn't be an incurable problem as many hams seem to think. This receiver has been designed to provide adequate decoupling from moderate high frequency noise components; however, severe cases may require external filtering. If you encounter noise problems in any radio in the car, the first thing to do is to disconnect the antenna to see if the noise is radiated noise. If you have this problem, good luck! You need to give your car a good tune up. Fm receivers are relatively immune to this type of noise.

If the noise persists with antenna disconnected, it is getting into the audio circuits through the power line. You should have the positive and negative power leads connected directly to the battery, since that is the only filter component in a vehicle, and the alternator puts out rectified ac without any other filtering than that provided by the battery. Ignition noise will also generally be present on all wiring in the car except directly across the battery.

If you have taken these preventive measures and the noise still rides through on power wiring, you should install an L-C filter in the positive lead of the power wiring. A choke in series (about 1 Henry) and a capacitor (about 1000 μF) from the radio side to ground should be adequate to filter out the ac component for all but extreme cases. If you also have a solid-state transmitter, the noise will probably also get into the transmitter audio; so you should install the filter in the battery lead to both receiver and transmitter while you are doing it.

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2. G. Francis Vogt, WA2GCF, "Speaker Driver Module," *ham radio*, September, 1972, page 24.

ham radio

performance of rf speech clippers

This article sheds new light on rf clippers, shows how to get maximum performance, and explains why results are often disappointing

Leslie A. Moxon, G6XN, Hampshire, England

For the last decade or so, spectacular claims have been made in regard to rf clipping of ssb signals. Despite many confusing, and often conflicting, statements in the literature, there seems to be general agreement that rf clipping can increase talk-power by at least as much as the average linear without infringing on license regulations or significantly affecting the speech quality; this is achieved for the cost of an extra crystal filter, plus a few diodes and resistors. These claims have not been rebutted and, when seeking reasons why the system has not come into general use you find much confused thinking, some unsolved design problems, and unexpected facts about existing transmissions!

I have used rf clipping exclusively for many years during which it has been embodied in a total of six homebrew transceivers, including several ultra-low power transistor rigs used for propagation experiments. These had to be ultra-portable for use on steep ground slopes, often without road or even footpath access, and the role of rf clipping was to increase average efficiency, reducing battery drain and weight as well as permitting the use of inexpensive output transistors. DX worked on these occasions, with less than two watts and often less than one watt PEP output, included

many contacts over the long path between Europe and Australia.

At the first attempt VK2NN *and* VK3IP were worked with about 0.5 watt. Later, a daily schedule was operated successfully with the same two stations from GM6XN/P during a six-day vacation on the Isle of Mull. During a retirement tour of Australia and New Zealand, eight European stations were worked in a single session from VK7LM, and reports up to S8 were obtained from England while operating ZM1BJF on Stewart Island. In one session, 13 U. S. stations were worked from G6XN/P on 21 and 28 MHz. There appeared to be little or no distortion attributable to the clipping and it is estimated that most of these contacts would have been impossible with an unclipped signal.

With efficient clipping it is easy to demonstrate an advantage of the order of two S-units by reducing gain to the point where the peak rf level (by then reached only infrequently) just starts to drop. Unfortunately, anyone expecting to gain all of these two S-units in practice is likely to be disappointed because the noclipping condition described above, though strictly correct, is unrealistic. It is now obvious from an analysis of amateur ssb signals that in most cases some overdriving of the final, which is then acting as a clipper, occurs for a small percentage of the time. It is a simple fact of life that as the drive increases, interference to adjacent channels increases, slowly at first, then more rapidly, whereas talkpower does the opposite.

The dividing line between *right* and *wrong* operation is not precisely definable and it would appear that by the time splatter becomes sufficiently continuous to give rise to complaints, there is already very considerable rf clipping taking place in the transmitter. On the other hand, if the drive is nearly constant (if it consists of an rf clipped signal or a two-tone test signal), any spillover into adjacent channels will be more or less continuous, and for any given average level of out-of-band radiation (however small), much greater care must be taken to avoid overdriving.

In view of all the subjective aspects there is some doubt as to the existence of a *solid engineering* basis for determining the correct adjustment of ssb transmitters with an unprocessed speech input. The use of clipping, which determines a definite peak level, has obvious value in this regard. Existing rf clipper designs using cascaded pairs of standard filters are, however, open to criticism on the

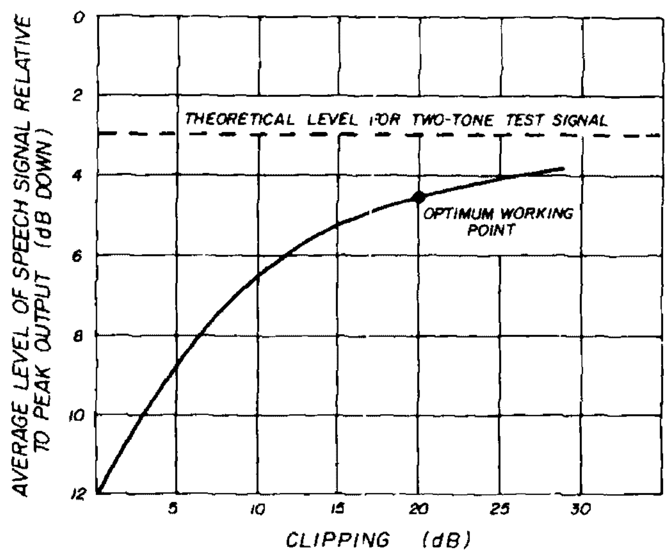

fig. 1. Variation of average signal output with clipping.

grounds of restriction of audio response, and an unnecessarily steep-sided overall response with corresponding inflation of manufacturing costs.

how it works

Rf clipping works in two different ways.[1] First, by providing an 8 or 9 dB increase in mean power level for a given peak power; secondly, by increasing the signal-to-noise ratio of speech sounds such as "th" which are important for intelligibility, but low in energy content, so they easily get submerged in the noise. This latter point, despite considerable discussion in reference 1, appears to have been ignored by later writers. It has been demonstrated[1,2] that, in *ordinary speech*, intelligence is conveyed entirely by frequency and not by amplitude. Unfortunately, large amplitude variations are

present so that, to prevent occasional peaks of unclipped voice from exceeding the transmitter peak-power rating or the license limit (whichever comes first), the mean power level has to be held down to about 12 dB less than the peaks.

If we clip the peaks of the ssb signal so that the maximum amplitude is reduced to some *small* fraction of its original value, the audio gain control can be advanced by the same amount so that the peak level is restored to its previous value. When most of the amplitude variations have been removed the mean level is, of not that practical clippers have a limited dynamic range.

Only a few harmonics in the modulation envelope actually get through the filter. Those that do get through cause some distortion and amplitude variations, but apparently the benefit of clipping before sideband-filtering is not completely neutralised, so even *audio* clipping remains effective to some extent.[3]

practical advantages

Fig. 1 shows that in one typical instance the mean rf power transmitted was increased 7.4 dB by 20-dB of clipping. This agrees reasonably well with the 8-dB increase found by other writers.[2,3,4] This is less than the average difference of two S-units in signal reports which I have obtained by reducing gain so that,in effect,you slide down the curve of fig. 1 from the 20-dB to the 0-dB point.

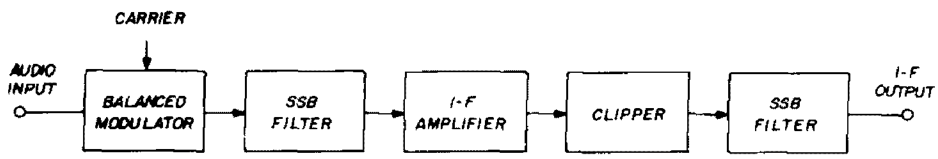

fig. 2. Essential rf clipper details.

course, much higher. **Fig. 1** shows how this mean level varies with the amount of clipping as I measured it, using my voice and rig.

The weaker speech sounds (when not accompanied by louder ones) will be boosted by the amount of clipping so long as the peak power limit of the transmitter is not exceeded.

Since it is essential for the clipping to be applied to the ssb signal, two filters are needed as shown in fig. 2. The second filter is required to remove out-of-band signals generated by the clipper; the role of the first filter is less obvious, but without it, spectacular things could happen at the output of the second filter. This is evident if you remember that the job of a clipper is to hold the *total signal* constant, whereas the sum of two sidebands is zero at one point in the cycle, the two voltages being equal and of opposite phase. Therefore, to keep the total voltage constant the voltage of each sideband has to become infinitely large; remove one to them so that the other is left unopposed, and the sparks fly! The output of the second filter, far from being constant, would consist of a series of sharp spikes. Spikes large enough to destroy vacuum tubes and circuits were it

The difference, perhaps, may be explained by threshold effects including the enhancement of weaker speech sounds, which has been investigated elsewhere for the case of 20-dB clipping. As would be expected, the gain from this cause was found to be closely linked to the prevailing level of word intelligibility; at the 70% level, which would probably be reported as QSA5, the total benefit from both effects was found to be 11 dB. At 50% intelligibility, corresponding to a higher noise level, it rose to 13 dB.

At very poor signal-to-noise ratios, useless for ragchewing, but probably better than nothing during a contest, or when trying to work a new country, the advantage drops again; this is presumably because the high-level components of speech, unaided, convey a certain small degree of intelligibility even though the weaker sounds may have been irretrievably lost.

From fig. 1 it appears that increasing clipping beyond 20 dB produces little further increase of mean level, and half the total enhancement is generated by a mere 6 dB clipping. This is consistent with reference 1. The linear relationship claimed by some authors ignores the rather spiky nature of speech waveforms; thus, for the first few dB of clipping the only energy lost is that of a few infrequent spikes so the mean level increases by almost the same amount. On the other hand, with 20 dB of clipping, the clipper operates well below the mean level of speech so the output is already nearly a square wave, and little further increase is possible.

The input-output relationship must be of the nonlinear form indicated by fig. 1. A similar result would be expected from allowing the final to overload except that in this case some of the transmitter power is radiated in adjacent channels where it causes interference. It is difficult to define precisely the point at which such interference becomes intolerable. However, legal requirements apart, it depends on many factors such as distance to the nearest neighboring ham and the quality of his receiver.

Interference caused by the first few dB of such clipping, compared with the next few, will be relatively small. For example, typical speech signals exceed 50% of the peak amplitude for only about 8% of the time. Therefore, if the final is allowed to clip the peaks by 6 dB, the adjacent channel interference will exceed that produced by a standard two-tone test for 8% of the time. International regulations allow a two-tone intermodulation products level of about 30 dB. In effect, an S9 signal is permitted to destroy an S4 signal on an adjacent channel.

In contrast, overloading by speech signals to the extent of the above example will cause some annoyance, but the weak signal will remain more or less intelligible because of gaps in the interference. As a fact-finding exercise fig. 3 shows the amount of "clipping" measured for a random sample of 60 amateur signals on the 14-MHz band. These were deduced by measuring the ratio of peak-to-mean signal level as described later. In obtaining this data no instances were observed of adjacent channel splatter. It seems likely that ALC systems, which normally have too long a time-constant to provide satisfactory compression,[3] nevertheless reduce the frequency and duration of overloads to acceptable limits.

Any method of increasing mean power obviously must increase the amount of interference generated by cross-modulation in neighbouring receivers and rf clipping is no exception. However,

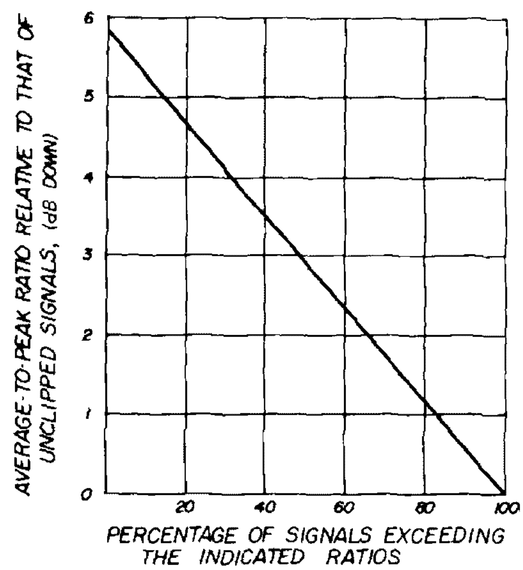

fig. 3. Average-to-peak power ratios for a typical sample of amateur ssb signals.

because of the highly nonlinear character of such effects it should be less damaging than a corresponding increase of mean power without restricting the peaks.

Most writers regard 20 dB as the optimum clipping level, and that agrees with my experience. It is possible that more clipping might be useful under very noisy conditions in order further to enhance the weaker speech sounds, but this has been difficult to prove or disprove. In general, there is a tendency with increased clipping for sibilants to become objectionable. Up to 20 dB of clipping, changes in voice characteristics appear to be small compared with other influences such as the microphone, filter response, etc., provided the system is operating

correctly. However, in the absence of noise a slight alteration in the voice has been observed.[1]

Despite the general agreement between writers on an increase in mean level of 8 dB for 20 dB of clipping, there are considerable differences in values quoted for peak-to-mean amplitude ratios of unclipped speech. Defining the "peak" as the level exceeded for 1% of the time, reference 1 finds a peak-to-mean ratio of 12 dB for an unclipped audio signal, and 10 dB for the modulation envelope of the same signal converted to ssb. Other writers quote ratios of 14-16 dB for the

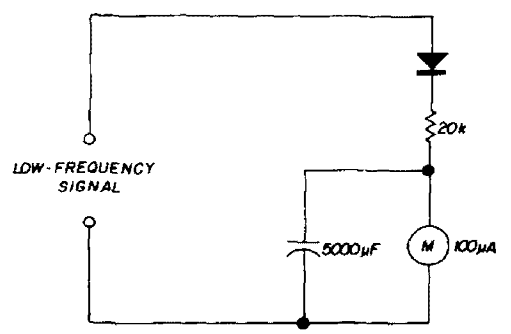

fig. 4. Long time constant, mean signal level indicator. Low-frequency signal should be 1 to 2 volts rms; resistance of meter should be approximately 700 ohms.

audio signal. Fig. 1 shows the ratio is 12 dB for the ssb signal but this could be in error by one dB or so as the zero level was difficult to estimate precisely. This ratio is unchanged if the average level of a single tone is retained as reference, but for consistency with other writers it's necessary to relate to the peak level of the single tone. That is, to add 4 dB so the 12-dB ratio becomes 16 dB.

The data for **fig. 3** were obtained using alc with a fast attack (20 ms) and slow release (4 s) time constant to hold the peak levels constant. The mean level was observed with a voltmeter time constant of 3.5 seconds (**fig. 4**). The values recorded were the highest maxima reached during continuous speech, but with some operators there were too many pauses and about 25% of observations had to be discarded as useless.

Zero clipping was taken as the lowest of the recorded values for steady talkers on the assumption that there must be some unclipped signals around. This was also confirmed by observations on one sideband of typical broadcast signals. The zero level, 12 dB down on a continuous tone, agreed with the rf measurements. The only signal known to have rf clipping (Comdel processor) was near the top end of the range. The average signal enhancement was found to be 3 dB, corresponding to 4.8 dB of clipping, or nearly half as effective as full rf clipping. Adjacent-channel splatter was not detectable but there were no local signals available.

Reference has already been made to the need for greater linearity in the final stage when clipping is used as any tendency to saturation will generate more splatter in view of the greater time spent at high levels. Other things being equal, the rf gain control cannot be advanced quite as far. In other words, for a given peak power capability, there is something to be subtracted from the increase of mean power obtainable by clipping. On the other hand, if the license specifies the maximum peak output power, and the transmitter is operating well within its rating, full benefit is obtainable.

This point does not seem to have been considered elsewhere although reference 1 suggests class-C operation of the output stage, and presents measurements of out-of-band radiation which were found acceptable provided the final was not driven beyond 90% of saturation. However, out-of-band radiation was increased, and it is not clear that there is any significant increase of efficiency by using this mode of operation.

So far, the discussion has been directed toward improving signal-to-noise ratios or extracting more intelligence with bad signal-to-noise ratios, but the same arguments should also be valid for QRM situations.

objections of rf clipping

Rf clipping has been slow in gaining support. There are several reasons for this, including prejudice engendered by audio clippers, much of it dating back to

the days of amplitude modulation. Despite distortion which takes the form of in-band harmonics, audio clipping is basically sound in the a-m case and unsound with ssb. However, in both cases the assumptions are undermined by practical considerations to the extent that some benefit is achieved from audio clipping in the ssb case. Disastrous consequences may result from its applicaton to practical a-m signals despite impressive performance in laboratory tests.

In a-m, the increase of modulation percentage due to clipping increases the importance of preserving the relative phases and amplitudes of carrier and side-bands to avoid distortion. Unfortunately, due to selective fading, these phases and amplitudes are more or less randomised, and clipping can reduce the intelligibility of an otherwise readable signal to zero. The effect is aggravated by agc which ensures that maximum distortion, caused by carrier fades, is accompanied by greatly increased volume.

These effects are not present with ssb, but frequency-distortion can still occur, causing undue emphasis of sibilants, and may account for occasional preferences for a lower clipping level when selective fading *is* present.

design considerations

The first few decibels of possible speech-processing gain are obtainable by simple methods such as alc, audio clipping or mere carelessness in adjusting the drive to the final. In contrast, the last few dB involve much larger amounts of clipping and this in turn requires careful design and monitoring. It should be apparent, from the earlier discussion, that clipping must take place only between the filters, the clipping level should be sharply defined, and subsequent amplifiers or modulators must have better-than-standard linearity at higher levels.

Furthermore, carrier suppression must be better than normal by an amount equal to the clipping; for example, with 20 dB of clipping, 20 dB of carrier suppression would add up to no carrier suppression at all during speech pauses. The second filter may provide the extra overall suppression needed, but this is not the whole story since the amount of carrier present at the clipper must be negligible compared with the clipping level. Otherwise, the beat note between the carrier and the wanted signal will appear as modulation.

To reduce this to 10% with 20 dB of clipping, 40 dB of carrier rejection is needed ahead of the clipper. This is only just within the lower specification limit for a typical IC balanced modulator such as the Plessey SL 640C plus a typical crystal filter. With 20 dB of clipping this still leaves the carrier only 20 dB below peak output, or about the same as it would be with no first filter or clipper. The second filter must provide the same amount of carrier-rejection as in a non-clipping modulator.

For maximum carrier rejection the balanced modulator should be operated at as high an audio input level as possible without risk of overloading; overloading at this point will not cause splatter or upset the action of the clipper, since out-of-band products are rejected by the first filter, but it will produce in-band, audible distortion.

Within the limits imposed by the above conditions, clipping level may be adjusted by alteration of audio gain or the i-f gain between balanced modulator and clipper. Reduction of i-f gain has the advantage of giving better carrier rejection. However, particular care must then be taken to ensure that there is no leakage of dsb signal which can bypass the clipper and filters if insufficient care is taken in layout and decoupling.

If the gain is increased too far, a point will be reached where the acoustic background noise in the shack fully modulates the transmitter; long before this point is reached the signal becomes unpleasant. The louder you speak and the quieter the shack, the more clipping can be used before this factor becomes important. Up to 10 dB of clipping is usually possible before there is any problem, and up to at least 20 dB is possible before it becomes serious. At 30 dB of clipping, there also

tends to be trouble with sibilants and other normally-weak sounds which become unpleasant due to overamplification. These disadvantages are not offset by any useful increase in talk-power.

The dynamic range of the clipper must obviously exceed the required amount of clipping, and at *least 26 dB* is advised. Within this range the output should be level within a small fraction of a dB since any variation tends to reduce the mean available level.

Pre-emphasis, up to as much as 6 dB per octave, is usually recommended with speech clipping, but the theory is obscure. The required amount is best found by experiment to suit the operator, microphone, degree of clipping and the carrier spacing.

Rf clipping can provide such advantages as higher efficiency and reduced splatter but the case for it usually relies on the assumption that without clipping the transmitter is peak-power limited, either by its design or because of license regulations. Until this point is reached, it is normally simpler and cheaper to increase drive than to add speech clipping. The transmitter must, of course, be capable of tolerating the increased mean power which results from clipping, a point which cannot be too strongly emphasized.

filters

Previous articles on rf clipping have specified pairs of standard filters, presumably because nothing better has been available. This is not likely to produce optimum performance because the tasks required of conventional filters, pre-clipping filters and post-clipping filters are all different. However, the requirements for carrier rejection are more or less identical in all three cases.

The basic task of the first filter is to prevent unwanted frequencies produced in or before the balanced modulator from getting through to the clipper at a high enough level to generate appreciable in-band distortion. With adequate carrier rejection, the main problem is getting rid of the unwanted sideband. Assuming, for example, 20 dB of sideband suppression, the unwanted sideband will cause 10% amplitude modulation of the wanted sideband, thereby restricting the power in the wanted signal for a given total peak signal by nearly 1 dB. The unwanted sideband is blocked by the second filter, but the desired sideband is amplitude modulated at the beat frequency and its harmonics.

Many of the resulting "sidebands of sidebands" will get through the second filter and add something to the general noise level, but the main consequence is loss of wanted signal power. Even if this is only a fraction of a dB, fractions add up, and it is only by taking care of all of them that most of the extra bonus theoretically obtainable from rf clipping can be realized.

In arriving at a specification for the sideband rejection of the second filter there is no clear distinction between right and wrong; 30 dB rejection of the unwanted sideband will reduce the power loss to 0.25 dB, which is fully acceptable, but even 10 dB is better than nothing, the potential gain from clipping being then reduced by 2.4 dB.

Rf clipping produces third-order intermodulation products in the passband. For a two-tone test signal with 20 dB of clipping the worst IM level is about −13 dB. This is an extreme case and products generated by speech signals appear to be of negligible importance provided they are at least 10 dB down.[4] Out-of-band IM will be roughly comparable with the in-band products and need to be reduced by filtering to a level which does not add appreciably to that generated in the rf stages of the transmitter. IM must be reduced from −13 to −30 dB to comply with international regulations, or −40 dB before an S9 +10 dB signal ceases to interfere with an S4 signal on an adjacent channel. Therefore, adjacent-channel rejection of − 27 dB is desirable.

The two filters have similar requirements, except that the second filter must suppress both of the adjacent channels whereas the first filter need only suppress

the unwanted sideband. Suppression prior to the clipper of high, unwanted audio frequencies is also necessary, and it is probably most convenient to do this in the first filter.

Why not use two standard filters? Typically, these have a 3-dB bandwidth of 2.0 to 2.2 kHz which gives an optimum compromise between loss of speech quality and generation of too much interference. If two such filters are cascaded for transmission, and the signal passes through another one or two similar filters for reception, intelligibility will suffer; moreover, the amount of passband ripple acceptable for a single-filter system is halved for a two-filter system.

In transmitters, the amount of out-of-band radiation is usually determined by nonlinearity of rf stages rather than inadequacy of filters, and the steep-sided responses of standard filters are dictated mainly by requirements for reception. In this case, with two filters available, the individual bandwidths can be increased and the selectivity reduced so that when the two are cascaded the overall response is unchanged.

By operating the second detector at the lowest possible signal level, even further relaxation of the specification becomes possible, as proved by the adequacy of receiver performance with surplus crystal filters. This means that filters which meet the transmitting requirement will be adequate for transceivers.

I am indebted to the Yakumo Tusin Company in Tokyo for providing filters modified to my own specification, the 6-dB bandwidth being increased from 2.2 to 2.5 kHz. The 60-dB bandwidth was thereby increased slightly but overall response of the pair of filters remains more than adequate. During initial discussion with the manufacturers, it appeared likely that the cost and physical size of the individual filters could be reduced by about 30% if there was sufficient demand.

conclusions

In summary, it appears that the first few dB of gain possible by rf clipping are currently being realized with no special circuitry or deliberate intent by allowing rf stages to overload for a small percentage of the time. This has been pointed out elsewhere and condemned as unethical,[5] but there is considerable evidence, theoretical and practical, for reducing the charge to one of poor engineering.

Formal rf clipping provides the basic engineering advantage of establishing a much more definite peak signal level and is capable of providing a total increase in talk power of about two S-units with no adverse consequences. This requires about 20 dB of clipping and observance of the following conditions:

1. Use of filters designed for the job, filters correctly designed for conventional systems being unsuitable.

2. Hard limiting with a dynamic range in excess of 20 dB.

3. Adequate precautions to ensure that *all* limiting takes place between the filters.

4. Adequate carrier rejection *prior to the clipper*, as well as overall.

5. The transmitter must have adequate *mean* power handling capability.

6. Operation of the final must be highly linear, and, in general, must have a higher peak power capability.

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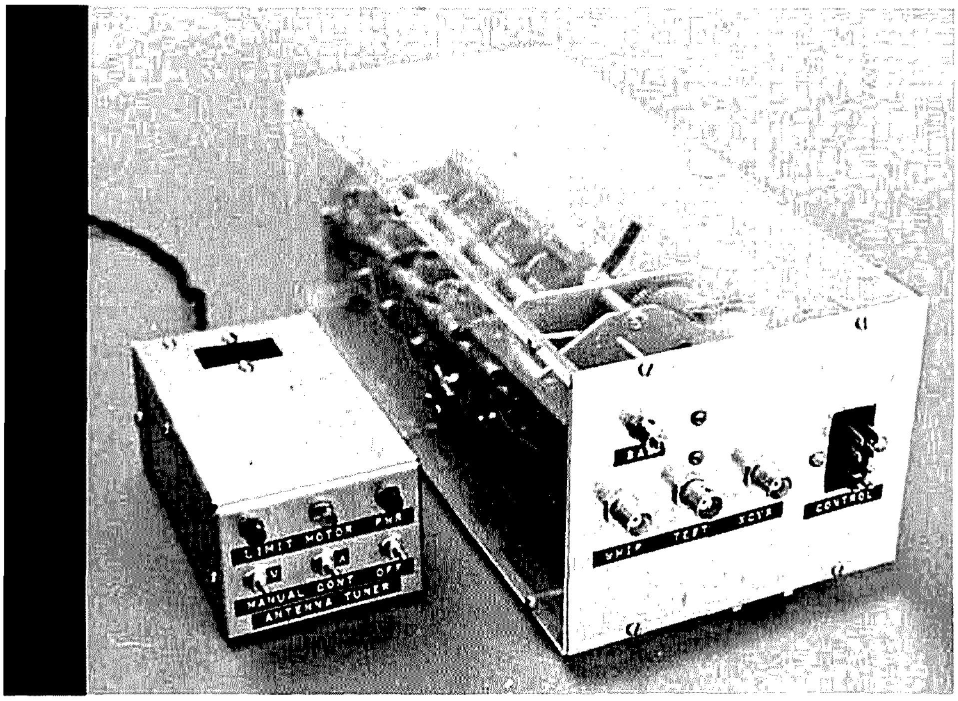
ham radio



automatic solid-state antenna tuner

A complete, automatic antenna-tuning system which is designed especially for the mobile operator

Stan Johnson, WAØAQC, 215 North River, North Aurora, Illinois

Of all the problems encountered by the mobile amateur radio operator, the antenna is the most grievous. Impedance mismatches and limited bandwidths on low frequencies can cause standing-wave ratios beyond the operating limits of most transceivers.

An antenna matching device[1] can eliminate the swr problem, but without a variable element on low frequencies, the narrow bandwidth remains. The tunable coil in series with the antenna[2] used by K6DY and W6WOY provides a wider range of usable frequencies, but watching the swr meter while tuning the coil and trying to keep the car out of the ditch is a job only for the courageous, well-coordinated ham who happens to have four eyes.

The automatic mobile antenna tuner built by WØIGP[3] provides a solution to the mobile antenna problem by automatically tuning the antenna with a variable inductor until it appears resistive. The tuner described here is similar to WØIGP's design except that solid-state devices have replaced the tubes and relays, and the variable inductor is replaced

by a variable capacitor. The resulting, improved swr makes the solid-state automatic antenna tuner a worthwhile addition to any mobile installation.

theory of operation

The automatic antenna tuner consists of a motor-driven variable capacitor connected in series with the mobile antenna and controlled by a phase detector. The antenna is adjusted to resonate at a frequency just below the desired frequency range. At the desired frequencies the resulting inductive reactance from the antenna is cancelled out by capacitive reactance from the variable capacitor (C1 in fig. 1), making the antenna appear resistive. Since the antenna is electrically longer *than a quarter* wavelength, the resistive component of its impedance increases, and provides a better match to 50-ohm coaxial feedline.

the phase detector

The phase detector, which is similar to the Foster-Seeley fm discriminator, determines resonance by comparing the voltage and current phase relationships at the base of the antenna. A detailed description of the phase detector theory of operation can be found in references 3 and 4.

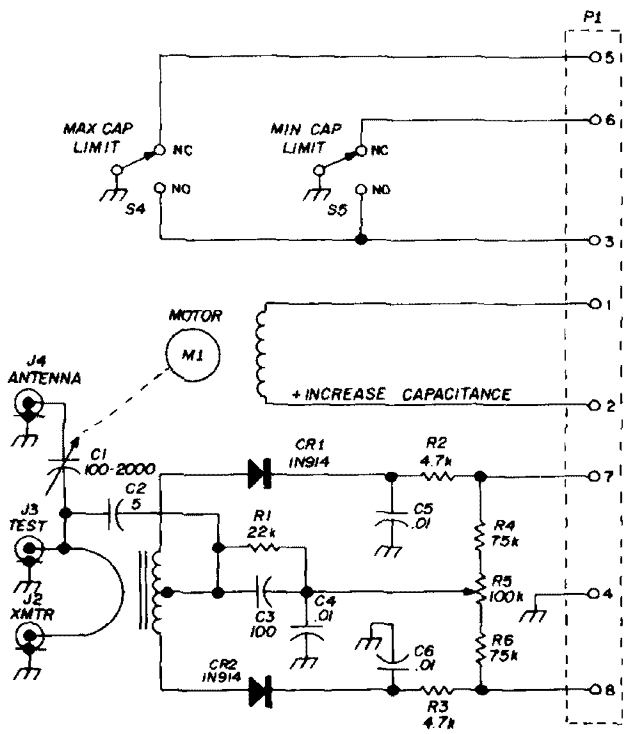
The center conductor of the coax from the transmitter is passed through a toroid to form the primary winding of T1 (see fig. 1). The induced voltage on the secondary of T1 is added to a reference voltage developed by the voltage divider, C2 and C3. The vector sum of the voltages is rectified and filtered to produce two opposing dc voltages on pins 7 and 8 of the output connector (P1). With a 50-ohm non-reactive load connected to test jack J3, and normal transmitter power applied, R5 should be adjusted for 0 volts between pins 7 and 8.

When the load is reactive, as in the case of an off-resonant antenna, a dc voltage will appear between pins 7 and 8, its polarity corresponding to the reactance of the load. During normal operation, the transmitter is connected to J2, the antenna to J4, and J3 is left open.

control unit

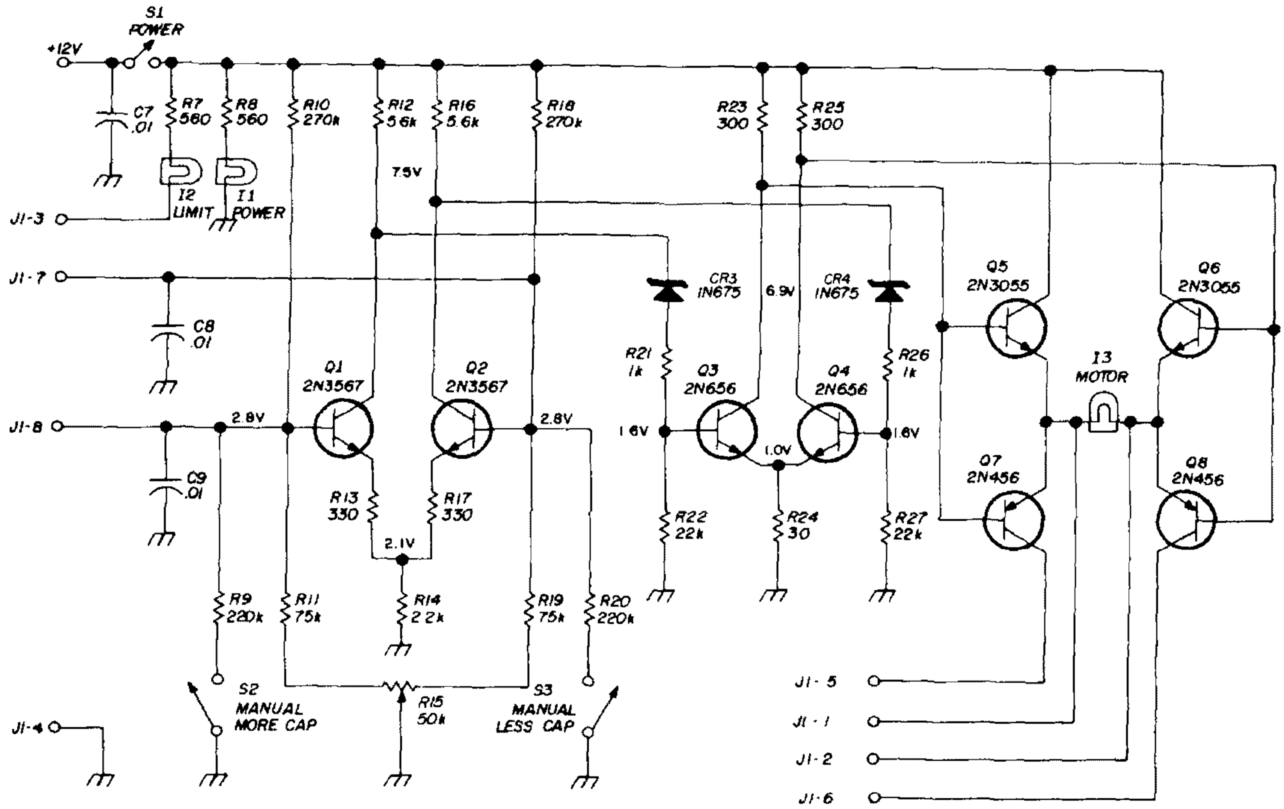
The dc voltage from the phase detector is applied to a differential amplifier, Q1 and Q2 (fig. 2). For manual operation, S2 or S3 can be used to unbalance the amplifier and tune the antenna. The tuning rate under manual control is determined by the value of R9 and R20 with maximum speed occurring at 90 kilohms.

Differential amplifier Q3 and Q4 further amplifies the error voltage to the level required by the motor-control



C1	100 to 2000 pf variable capacitor with a 120 to 1 gear reduction
M1	24-Vdc motor manufactured by Globe Industries, Inc., Dayton, Ohio (part number B3A1225)
S4,S5	spdt microswitch with leaf
T1	Toroid core, E type (red color code, 0.5 to 30 MHz) 0.25" ID, 0.45" OD, 0.14" high. Obtained from Tri-Rio Electronics, 2614 Lake Shore Drive, La Crosse, 2, Wisc. 54601. Primary consists of coax center conductor passing through toroid core. Secondary is 2 wires, 24 gauge, enameled, bifilar wound for 9 turns, connected to provide a center tap.

fig. 1. Tuning unit containing motor, tuning capacitor, limit switches and phase detector. Control unit is shown in fig. 2.



I1,I2,I3	15 mA, 3V lamps
Q1,Q2	matched pair of 2N3567s or similar transistors with current gain of approximately 75
Q3,Q4	matched pair of 2N656s
Q5,Q6	2N3055 or similar
Q7,Q8	2N456 or similar

fig. 2. Schematic diagram of control unit. It is connected to tuning unit (fig. 1) through an 8-conductor cable.

bridge circuit, Q5, Q6, Q7 and Q8. The selection of the control-bridge circuit transistors is not critical. Any transistors, either silicon or germanium, capable of handling the motor current should suffice.

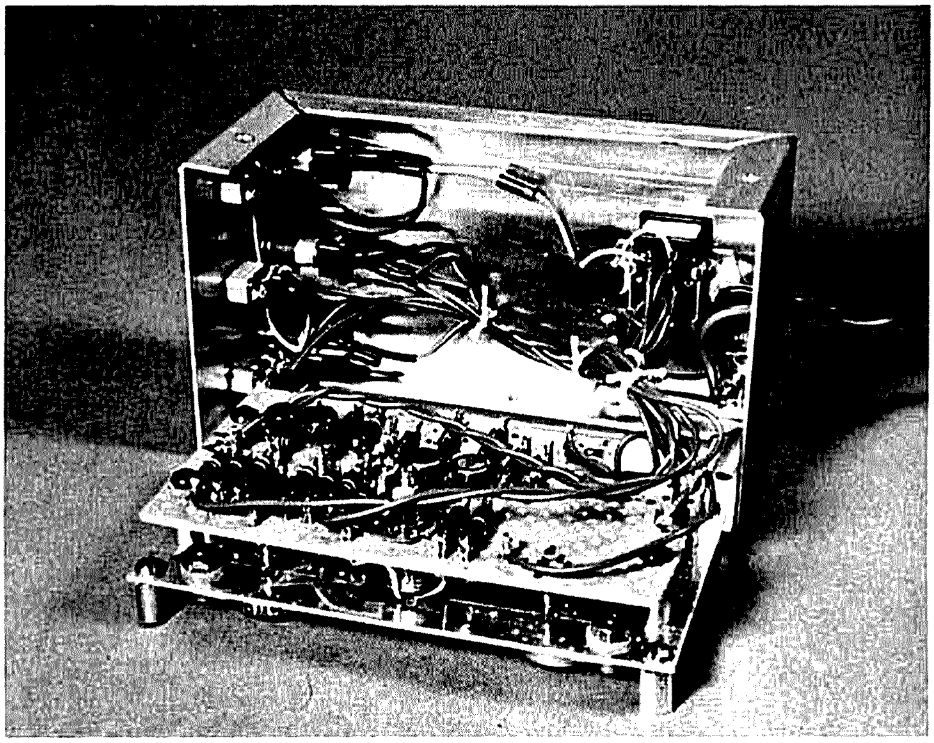
When the motor-control bridge is unbalanced, either Q5 and Q8 or Q6 and Q7 will conduct, passing current through the motor and rotating C1. If C1 has reached the limits of its rotation, limit switches are actuated, removing ground from the collector of Q7 or Q8 preventing further rotation in that direction. The control unit has three indicator lamps, I1 for power, I2 for when C1 has reached the end of its rotation, and I3 for motor voltage.

For best performance, Q1 and Q2, CR3 and CR4, and Q3 and Q4 should be matched pairs. To compensate for component differences, R15 is adjusted for zero volts between the collectors of Q3 and Q4 with no rf applied to the phase detector.

antenna tuner

Both the tuning motor and variable capacitor C1 are refugees from my junk box. A 24-Vdc motor was used. However, it does a very nice job of turning the capacitor on as little as 3 Vdc at 300 mA. The tuning capacitor from an ARN6 radio compass receiver is used for C1, but any reasonable facsimile should work.

An insulated gear is used between the motor and C1 to keep rf off the motor. Unfortunately, this gear was found huddling in a corner of the junk box without the foggiest recollection of its ancestry. Hopefully, the photographs will help identify duplicates. If an insulated gear cannot be located, perhaps an insulated coupling could be substituted. The tuning unit is housed in a plexiglass box with the limit switches attached to the sides. The

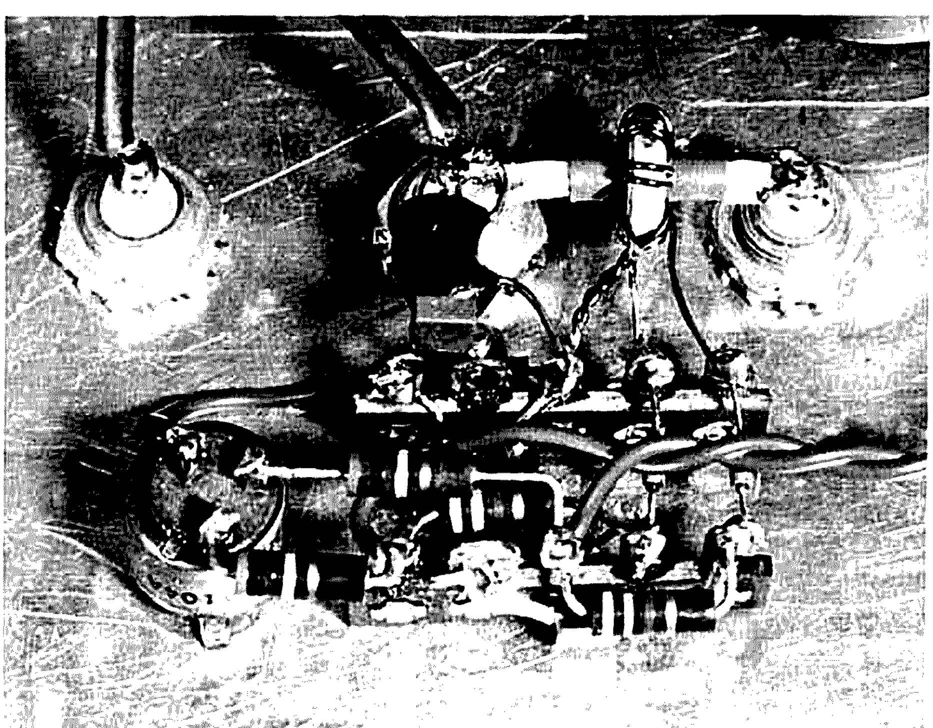


Control unit is mounted in 2" x 3" x 5¼" box.

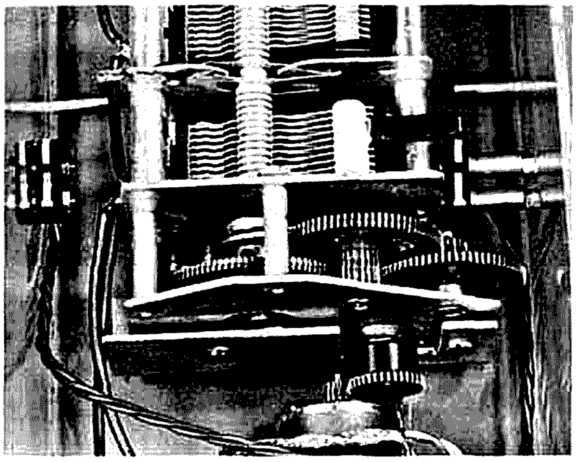
limit switches are actuated by an insulated metal strap which has been glued to the rotor of C1.

The direction of current through the motor windings determines the direction of rotation; therefore, it also determines whether C1 is rotating towards maximum or minimum capacitance. With one motor lead grounded, and voltage applied to the other lead, note the direction of rotation. The wire from the motor which, when made positive, increases the capacitance, should be connected to Q6 and Q8 via pin 2 of J1.

The other lead will cause the wrong limit switch to be activated when the capacitor reaches its rotation limit. The result will be wailing and gnashing of teeth. If the motor rotates in the wrong direction when rf is applied to the phase



Construction of phase detector in the tuning unit.



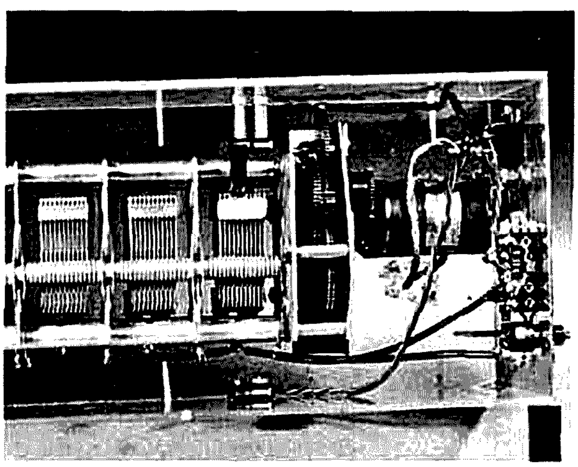
Tuning unit closeup showing limit switches and actuater.

detector, simply reverse the wires on pins 7 and 8 of P1.

The tuning unit in the car trunk is connected to the control unit with an 8-conductor rotor cable. It is recommended that the two larger wires be used for the motor leads (pins 1 and 2).

results

After installing the antenna tuner in the car the swr on 10, 15 and 20 meters was negligible. On 40 meters, the swr at 7250 kHz was 1.0:1 and rose to 1.5:1 at the edges of the phone band. Without the tuner, on 75 meters the bandwidth (swr of 2.0:1 or less) was 16 kHz, while with the tuner, the bandwidth equaled 84 kHz. On 75 meters the bandwidth is limited on the low end by the antenna's resonant frequency and on the high end by the increasingly large resistive component of the antenna's impedance.

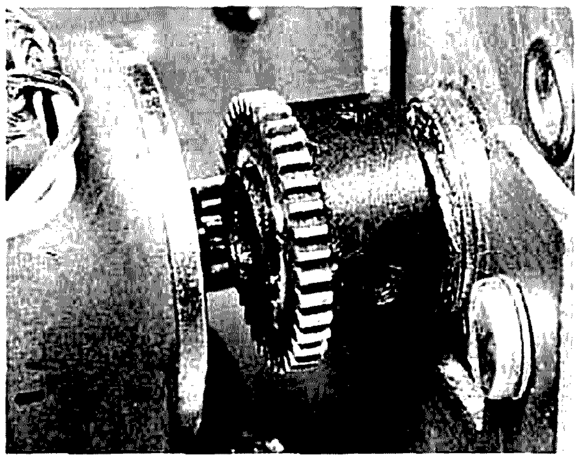


Inside the tuning unit. Note the small strap which is glued to the rotor to actuate the limit switches.

The antenna tuner also eliminated the effects of heat dissipated in the resonator which changed the resonant frequency. The most pleasant benefit derived from the tuner, however, is that the transmitter always sees a resistive load, and therefore does not require retuning after large frequency excursions. After the transmitter frequency has been changed, a short gleeful whistle into the microphone is all that's required to tune the antenna to the new frequency.

credits

I would like to express my appreciation to my friends at Bell Telephone Laboratories for their help; Bill Thelen for conversations on differential amplifiers, Jon Fistler for the photographic work, and Ron Cunningham, W9MAF, for editing assistance.



Insulated gear

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ham radio

5-MHz WWV receiver

High-performance frequency-calibration receiver uses new reciprocating detector circuit for maximum frequency stability and accuracy

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Since the day of the "Lunch Box Frequency Meter"[1] has long since passed, the FCC frequency-measuring requirement is now met through the use of a crystal-controlled frequency standard, with appropriate marker outputs to help indicate band edges. These standards have increased in popularity to the point where most receiver manufacturers include one as part of the receiver.

A lot of interest has been directed to the accuracy of frequency measurement; this activity is due to the national contests and the MARS programs which require nets to operate within confines not previously required of amateurs. Most amateurs use the station receiver to calibrate their marker generators. few have a separate receiver for this purpose. Very few have a receiver which can be used as a "sole-source direct calibration signal."

It is the intent of this article to describe a receiver which is primarily for this purpose. It is a simple, but very effective, self contained, battery-operated solid-state unit to which a frequency standard may be calibrated very accurately.

This receiver has the single purpose of receiving WWV at 5 MHz, but it could be put on any other frequency which you want. To change the operating frequency all that is required for the most part will be a new set of coils.

circuit

The rf circuit consists of two tuned rf amplifiers and a synchronous detector. A trf configuration was chosen because the receiver is not dependent upon any frequency-determining devices other than the tuned circuits in the rf amplifiers; this eliminates errors in frequency calibration. Since the output frequency is the operating frequency it is dependent upon the received signal for its accuracy.

The receiver circuit shown in fig. 1 can be used on many frequencies. At W1SNN it has been built for both 60 kHz and 5 MHz. At WA1NWF it has been duplicated for 10 MHz where it also used as an i-f strip for a 50-MHz receiver.

To change the inductors to new frequency ranges, information on how to wind toroidal inductors can be found in articles published in recent issues of *ham radio magazine.*[2,3,4]

The input circuits of the receiver may look a little *strange*, but they are built to accomodate an antenna and the output of a calibrator. Each input is isolated from the other — the reason for their existance will be described later under operation.

The first and second rf stages have tuned input and output circuits which are very loosely coupled. The loose coupling minimizes loading the toroids down and lowering the Q of the tuned circuits. The unloaded Q of each coil was 180; calculations indicate that the loaded Q is in the vicinity of 110. By loosely coupling the output of the first rf stage and the input of the rf stage preceeding the detector, an effect of critical coupling is possible which enhances the detection bandwidth. This improvement in selectivity is very necessary. Operators who think that 5 MHz is a sacred frequency used only by WWV will be chagrined to find that adjacent signals slop over from as much as 20 kHz away.

Fets are used to further preserve the high Q of the tuned circuits. The fet inputs are lightly coupled to the tuned circuits.

To insure circuit stability each rf amplifier is neutralized. The neutralization provides stabilizing feedback and insures spurious oscillations will not be present. Further, each stage is completely shielded. Seems like a lot of work? Not by a long shot; when you turn the receiver on, it all stays in the box and does not oscillate.

detector

The reciprocating detector circuit was previously described in *ham radio*.[5] This circuit has several advantages not found in other detectors. The detector is synchronous and depends upon a reference signal which is synthesized from the received signal. The reference signal is proportional to the average received signal; therefore, it is always at just the right level, eliminating noise which is generated by conventional reference oscillators.

The reference signal is synthesized through a narrowband rf filter. Because of the Q of the filter, it will not respond to impulse noises. This eliminates interference caused by short duration impulses, such as static crashes. Selective fading which is common on double-sideband transmissions is reduced, and in many cases, corrected with this detector. In the form presented here, the detector will respond to any of the modes used by amateurs (except fm) without any circuit changes. Last, but most important, this circuit provides the reference signal for calibration.

construction

The two rf stages and detector were built on pieces of copper-clad epoxy board. Each circuit was built on a separate board as I am experimenting with the circuits in other applications. A single piece of circuit board can be used. In my unit the boards are 2¼ by 4¼ inches; components were mounted with a simple technique described by W6CMQ.[6]

W6CMQ used a fly cutter to cut rings in the copper, providing an insulated tab for part mounting. This method provides an excellent mounting for small parts and semiconductors. A variation in this approach is accomplished by drilling small pilot holes so a Vector Pad cutting tool can be used. Vector terminals similar to *Flea Clips* can be inserted into the pilot holes. It is possible to mount components on the opposite side of the board which acts as a support for the coil shields; this is very effective for mounting the neutralizing capacitors.

The detector circuits and part of the audio system were mounted on a separate board, which is a printed circuit. All components in this unit were shielded with shields made from small pieces of sheet brass. The small subassemblies were fastened to a larger aluminum sheet which is the cover sheet for a chassis; controls and batteries were mounted in the chassis, which is inverted.

filter

The reciprocating detector includes a few components which need more discussion regarding their adjustment and construction. These components are found in the rf filter used to provide the reference

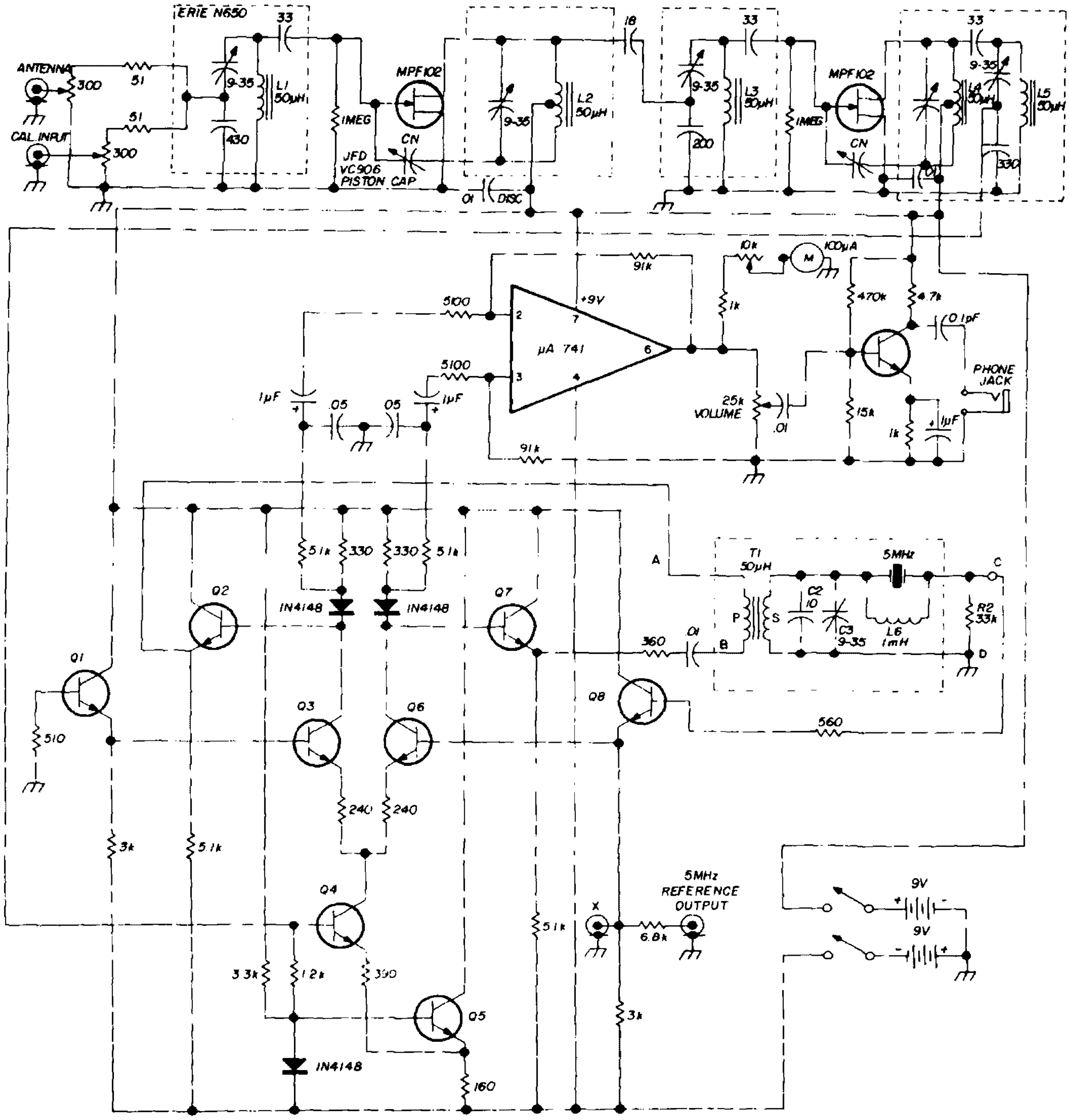


fig. 1. Circuit diagram for the 5-MHz frequency-calibration receiver.

signal from the received signal. The reference filter is narrowband, 500-Hz wide at the 3-dB points. It is not unlike the crystal filters used in early communication receivers. However, those filters were adjustable and did not attempt to maintain a uniform shape factor. *More often* they used the slope of the filter as a rejection point for an offending signal.

This filter has similar circuitry, but it is used to pass a signal. Therefore, the crystal must *not* be an old FT243 type, which has a high holder capacitance. In fact, almost all of the older free-plate type holders will give trouble when used in this circuit. An HC6/U crystal is very small and has about 0.5 pF holder capacitance.

To remove the loading effect which this capacitance causes, the crystal must have a parallel inductance to resonate out the holder capacitance. This is labeled L6

in fig. 2. The inductance is quite high, but is easily made with a toroid core.

The input transformer for the filter is shown in fig. 2 with the two windings marked P and S. Since the filter is driven differentially, it is convenient to use a coupling loop for the input circuit. However, the output is single ended. To wind the toroid properly, make the S winding first, starting around the toroid core using up about 330 degrees of the winding surface. Then wind P into the remaining 30 degrees. Enough coupling exists to provide transformation, but it will not add additional stray capacitance to the secondary winding.

The filter has low insertion loss. A loop gain of about three can be obtained using no amplification other than that of the basic reciprocating detector circuit. To prevent loading the filter impedance should be, and is, 360 ohms, by virtue of the resistor which is connected in series with the filter and the emitter of Q7. Increasing this resistance broadens filter response and reduces its noise rejection capability.

I have only discussed crystal filters for the detector. If a low-frequency circuit is required, sufficient Q can be obtained by using pot cores.

Fig. 2 gives three frequencies in decades. By scaling these, inductances can be determined for any frequency in these ranges. By using an appropriate toroid core and crystal it should be no problem for you to design your own reciprocating detector filter. A source of filters for those of you who do not care to make your own is available from me upon request.

The choice of transistors and ICs is left to the builder. In my case I used 2N3415 transistors and 1N252 diodes because these components were readily available. The IC is a Fairchild μA741 but I originally used a Motorola MC1433G. Any of the op amps of this variety can be used as can any of the silicon transistors and diodes of similar operating characteristics. Motorola MPF102 fets were used in the rf amplifiers because I had them, but I am sure there are many others in the same price range which can be used.

The 5-MHz rf output should be brought out as directly as possible from the printed-circuit board. Do not use shielded leads to the output jack for this purpose. Plan to locate the output jack on the adjacent to the side of the board so the 6.8k resistor can be connected directly to the output jack.

The meter is a center-scale 100 microammeter which does not indicate very much until you get right down to the nitty gritty of a frequency count. A larger movement will work — it just will not swing as far. The audio amplifier is sufficient to knock the phones off your head, but is not good enough to run a speaker.

Once you have the receiver built, there are several adjustments you have to make before you can use it as a frequency calibrator. First apply the ±9-volt supplies to the reciprocating detector and audio amplifier. Through a .01 capacitor apply a 5-MHz signal to the junction point of the trimmer and 330-pF capacitor connected to L5; the signal should be modulated. With the phones plugged in, adjust the secondary of T1 until the *modulation* is heard.

If a heterodyne is heard, it is an indication that the signal generator is not exactly on 5 MHz; it should be readjusted until you have a zero beat. To make this adjustment, turn off the modulation to the signal generator, set the sensitivity control full on for the 100-microamp meter. If the input frequency is off far enough to hear a beat signal, the meter will probably not indicate, or will quiver. Now adjust the signal generator until a zero beat is obtained. When you get close to zero beat, the meter will swing slowly around its zero center; reduce the sensitivity control to keep it on scale.

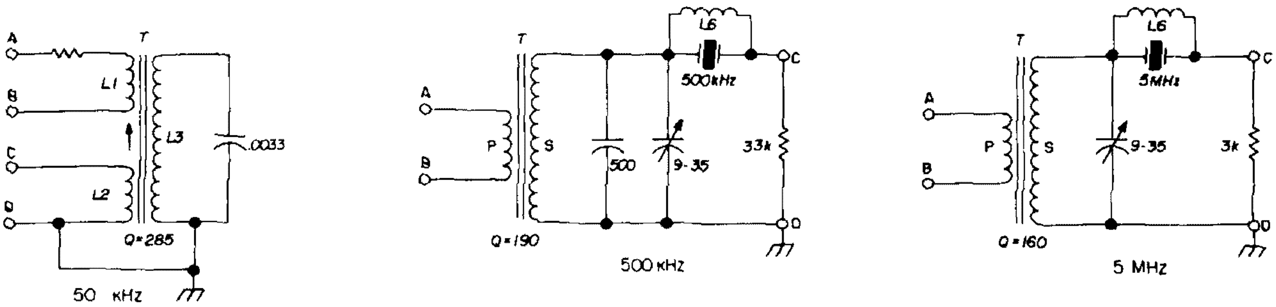
The signal generator will probably not be stable enough for further adjustment. (If the components used are those recommended there will be none required.) WWV will have to be used for the final adjustments.

Disconnect the signal generator from the detector circuit and connect it to the

antenna input. Modulate the signal generator; be sure the antenna potentiometer is full on. With a moderately high output from the generator the 9-35-pF trimmer on L5 should be adjusted until the modulation increases in amplitude in the phones. Once this L5 has been resonated, the trimmers on L4 through L1 should be adjusted for maximum signal. Decrease *signal volume by turning* the antenna potentiometer toward *off*.

By the time you have gone this far the capacitor for the first rf amplifier to a null, or until there is no indicated signal.

Remove the signal generator and rf probe; connect an antenna to the input jack and a crystal calibrator to the calibrator input. Be sure the antenna potentiometer is full on and the calibrator pot is near off. Do not turn on the calibrator yet; apply the voltage to the receiver. If you have tuned the receiver correctly to 5 MHz, you will hear the 5-MHz WWV signal. Observe the signal for a while and



50-kHz filters

L1 5 turns no. 32 enameled

L2 43 turns no. 32 enameled

L3 109 turns no. 32 enameled, 2.9 μH

All wound on Ferroxcube pot core, no. 1811CA250-3B7

500-kHz filters

T1 primary consists of 30 turns no. 32 enameled; secondary is 115 turns no. 32 enameled on Micrometals T44-15 toroid core.

L6 is 10 mH toroid

5-MHz filter

T1 primary consists of 10 turns no. 32 enameled; secondary is 96 turns no. 32 enameled on Micrometals T50-2 toroid core.

fig. 2. Filter table for the calibration receiver's reciprocating detector gives component values for different frequency ranges. For frequencies other than 50 kHz, 500 kHz or 5 MHz, values may be scaled as described in text.

receiver will have broken into oscillation. Turn off the power, connect the signal generator back to the capacitor junction point on L5, then to the junction point of the 9-35-pF and 200-pF capacitors connected to L3, and connect an rf probe to your vtvm.

Adjust CN, the neutralizing capacitor, until the signal goes through a null or disappears; move the signal generator to the same junction point where the vtvm rf probe was located and connect the signal generator to the antenna input. *Adjust the antenna potentiometer* to full on. Increase the output of the signal generator and adjust the neutralizing note if there are any beats or heterodynes. If there are, the rf stages may still be oscillating. If all is well, when the audio tones disappear from WWV, you should hear *nothing* except the timing ticks. Now turn on your crystal calibrator and adjust the calibrator potentiometer to half value. If your calibrator is right on frequency, or very close to zero beat, you will hear nothing.

Careful listening will reveal a very slow beat signal, which will have a hissing sound as it goes through zero beat. Adjust the two input potentiometers alternately until the signal levels have the same strength and the slow beat note becomes

more pronounced. All of this must be done during the tone-off period of WWV. Now turn up the sensitivity of the meter circuit; the meter will move very slowly through the center scale as it follows the beat, and as you adjust your calibrator trimmer, you can make the meter sweep faster or slower.

The idea is to get the meter to stand still at zero. It will probably not stay still unless you have a pretty good frequency standard, but you can listen to the time ticks from WWV and count the time it takes for a very slow excursion across the meter zero. You are now able to measure quite accurately the parts of a cycle per second at 1 MHz for which your calibrator is stable.

Now you can try the 5-MHz output reference signal. Turn off your crystal calibrator, and connect the output reference jack from the WWV receiver to the vertical plates of an oscilloscope. Connect the output of your calibrator crystal. If it is a 5-MHz crystal oscillator when the gain pots are set correctly for the oscilloscope defection plates a perfect circle will result. The circle will be stationary if the calibrator is right on, but will rotate slowly if the calibrator is slightly off frequency. The speed with which the display rotates can be timed in the same manner as described for the meter. If the calibrator crystal is 1 MHz, a chain-like Lisajou pattern will result; it will be stationary if on frequency.

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ham radio

first steps to satellite communication

Background, definitions and suggestions for entering the fascinating field of amateur satellite communication

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If we, as radio amateurs, are going to involve ourselves in satellite communications, we must develop what professional engineers call systems engineering. We must have a systems engineering attitude and approach to the problems of communicating via amateur satellite. Most radio amateurs think in terms of individual components and devices. There is no longer room for this kind of technical narrowness; for satellite communications are far more demanding than the typical twenty-meter single sideband contact.

sub-systems

First steps to satellite communications involve the following sub-systems:

1. A transmitter with an output power of 25 watts in the CW mode of operation is thoroughly adequate. Though it's hard to believe, anything more than that may overload the satellite's circuitry or cause interference with adjacent channels.

2. A communications receiver with good sensitivity, selectivity and stability, and a converter to receive all of the satellite frequencies — vhf and uhf.

3. An antenna array with a forward gain of at least 10 dB and, ideally, vertical *and* horizontal directional control.

Systems engineering simply means coordinating these three sub-systems into one workable satellite communications system.

transmitting sub-system

There is nothing really very special about the transmitter requirement for the 144 to 148 MHz and 420 to 450 MHz bands. Any standard circuit in present day use should work well.

From the standpoint of design efficiency — and again, systems engineering — we should coordinate each stage to interact interdependently. This means we should use the same oscillator, doubler, control circuits and power supply for both bands. When we are using the 144 to 148 MHz circuits all other irrelevant circuits are off.

Special care is necessary to avoid certain characteristic problems. Primarily, we need proper shielding and efficient matching of the impedance of the final amplifier to the transmission line. Some radio amateurs don't like to admit it, but this is a very frequent cause of failure.

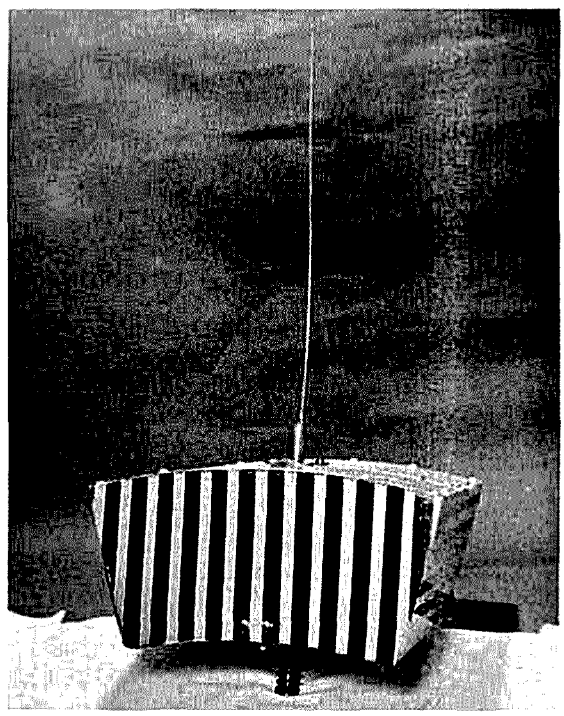
receiving

No attempt is usually made by amateurs with an interest in satellite communications to build from scratch a complete communications receiving system for vhf or uhf operation. It's very much more complex than building a transmitter and is not really necessary. Using a good quality dual-conversion superheterodyne communications receiver in conjunction with a low-noise, high-gain converter should achieve the same results. The converter's intermediate frequency output is usually either the 20- or 10-meter band.

antenna array

Using commercial equipment, antenna arrays are considerably simpler than any other sub-system. This doesn't mean an antenna isn't important — it's enormously important; but as long as we properly match the transmission line to the antenna, and the antenna's resonant frequency is close enough to the operating frequency, we should experience little difficulty.

Gain isn't necessarily important, though entirely desirable, to receive satellite telemetry transmissions. Indeed, results have shown that a vertical array with unity gain often works satisfactorily. Ground planes work. This, of course, is only for satellite telemetry reception. For reasonably reliable satellite two-way communications, with the satellite as an active relaying point in outer space, a directional array with about 10-dB gain is necessary. Vertical and horizontal directional control with computing devices to track the satellite are nice; but mounting a beam at a 25 to 45 degree angle on a television-type mast and rotator should work well much of the time. A little experimentation should tell us the optimum angle of radiation.



One of the early amateur satellites — OSCAR 2 — launched June 2, 1962. (Photo courtesy ARRL.)

Helical and parabolic arrays are extremely efficient but cumbersome to build and to operate. Yagi arrays with both vertical and horizontal elements have the advantage of switching from vertical to horizontal to circular polariza-

tion at will when using electrical remote-control coaxial switches. An antenna with an ability to change polarization is much more valuable than any other kind for amateur satellite service.

system losses

Critically important to satellite communications, the type of transmission cable and transmit-receive switch can easily determine success or failure. This and all the other variables affecting the strength of the signal, and subsequent success or failure, can be written down in what radio propagation engineers call the system loss formula.[1] This formula takes into consideration eight variables and is expressed LS = Lta + Ltp − Gt + Lp + Lrp − Gr + Lra + Laa, where LS is the total system loss,

Lta is the ohmic loss in the transmitting antenna and its transmission line, in dB,

Ltp is the polarization mismatch loss of the transmitting antenna,

Gt is the gain of the transmitting antenna in the desirable direction,

Lp is the path loss in dB,

Lrp is the polarization mismatch loss of the receiving antenna,

Gr is the gain of the receiving antenna in the desirable direction,

Lra is the ohmic loss in the receiving antenna and its transmission line, and

Laa is the antenna aperture loss.

ohmic loss

Ohmic loss in a transmission cable is the resistance the inner conductor and dielectric offers to radio frequency movement.[2] It is measured in decibels per hundred feet. On vhf, non-pressurized Alumifoam coaxial cable is applicable for this kind of service, and is made up of a low-loss polyethylene dielectric and seamless aluminum outer conductor, with a jacket of xelon polyethylene.[3] We might pressurize the cable for uhf applications under certain critical circumstances. Certainly, any coaxial cable should be in perfect physical condition. Coaxial cable installation is critical at any radio frequency for this kind of service.

A transmit-receive switch was mentioned operating under the assumption that the same antenna used for transmitting is also used for receiving. Silver-plating the contacts and careful construction will cut down the insertion line-loss, which is unavoidable.

One of the more sophisticated club stations designed for moonbounce and satellite communication. Complex equipment, however, is not that much of a necessity. (Photo courtesy Talcott Mountain UHF Society and the ARRL.)

polarization mismatch

When a radio signal is transmitted from earth to a satellite and back the polarization may change many times. Leaving the antenna, the radio signal may be of a vertical polarization, but after going through earth's atmosphere, through the satellite's amplifier circuits and then re-entering earth's atmosphere, the signal may change to a horizontal polarization — or most likely something in between. If the receiving signal is of a different polarization than that of the antenna, a polarization mismatch occurs causing a further reduction in signal strength. It could be the deciding variable, so it's important for us to know the receiving polarization and to have the ability to properly match polarization with the right antenna.

gain

Some antenna arrays will in effect amplify radio frequency power in certain directions. The primary concern of the system loss formula is to know the amount of rf gain there is in a particular array in the desirable direction of transmission, relative to a half-wavelength dipole or that hypothetical antenna, the isotropic source. The unit of measurement of gain is the decibel.

antenna aperture loss

Only a certain amount of radio frequency power coming from a directional antenna array will be of use in the desirable direction of transmission. The remaining power, if the antenna array is efficient, will be relatively small. Antenna aperture loss is a measure of that percent of the total available power delivered to the antenna which is wasted in radiation in directions other than the one chosen, focused beam toward the desired target antenna. This loss is measured in decibels.

path loss

Path loss is the total attenuation of the medium between the transmitting station on earth to the satellite in space, and between the satellite in space to the receiving station on earth. Path loss is a function of the medium characteristics and radio frequency energy in time. The unit of measurement is again the decibel.

At present, amateurs are mostly using 144 to 148 MHz and 420 to 450 MHz as their operating frequency range — and getting good results. The 28-MHz band is also in use but not as efficiently as the other two, though it was a surprise to learn that "10-meter signals from a satellite can penetrate the ionosphere at all elevation angles without experiencing blockage due to ionospheric reflection at low elevation angles."[5]

The favorite band is two meters because of the relative simplicity in the design and operation of the tank circuits and shielding. Also, reasonably priced tubes and transistors operate well there. An antenna system above the two-meter band must be larger to cover the same captivity area. Transmission cable is more critical above 300 MHz and hence gives unacceptable attenuation loss figures. This can be overcome, but only with a bigger diameter cable and a larger pocketbook.

acknowledgements

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ham radio

carrier-operated relay

Robert C. Heptig, KØPHF, and Robert D. Shriner, WAØUZO

Simplicity, dependability and an end to haywire lashups are benefits of this little gem

When repeaters first started to become popular a few years ago, amateurs desiring to build a repeater started to wire up what is known as a carrier operated relay. However, the words carrier operated relay are actually a misnomer. All repeater designers apply the voltage derived from the squelch or the limiter to the circuitry that operates the relay. This really doesn't make any difference, as all we are after is a relay action caused by the presence of a signal into the receiver.

After constructing a COR that will key the transmitter, a time-out timer should be constructed to limit the time that the transmitter can be on the air. From one to three minutes is a common time for the time-out function.

Too often we have a bucketfull of garbage haywired into the repeater consisting of one relay operated by the incoming signal which in turn operates the relay that keys the transmitter that is released by yet another relay that is controlled by the time-out function. This is all shown in fig. 1.

A better concept of how to control a repeater is shown in fig. 2. It is obvious that the reliability will be greatly increased due to its simplicity.

Other uses for this COR can quickly be seen — such as for a remote tape log in which case the value of C3 and R13 can be changed to limit the length of time that the tape log runs on each trans-

Etched and drilled G10 epoxy circuit boards are available from Circuit Board Specialists, 3011 Norwich Avenue, Pueblo, Colorado 81008. The price is $4.50, postpaid.

mission, (10 to 15 seconds being a normal time). Another use that is becoming very popular is to connect the relay to the speaker of base or mobile receivers and adjusting the time to about 10 seconds. Then all you will hear is the first few words of a transmission, and if the call isn't for you, then you don't have to hear it. If you do want to listen, though, the feature can be disabled.

circuitry

Practically any voltage and any relay can be tied to the circuit and it will go right to work. Primary power at 12 Vdc will be assumed for this discussion along with a 12-volt relay. However, keep in mind that the circuit will operate on any voltage from 6 to 36 Vdc. Almost any relay will work and the relay need not be the same voltage as the circuit uses.

The input signal to the circuit must be a negative voltage. This is common in most tube-type receivers and can be found in the squelch or limiter area of the receiver.

If the point that you desire to tie into in your receiver goes positive when receiving a signal, then simply replace Q1 with a P-channel field effect transistor such as a HEP 803 and reverse the zener diode CR1.

Now, to get down to the nitty gritty of the circuit. A negative voltage applied to the input is fed to Q1 by way of R1 which acts as a current limiter and prevents any loading of the receiver. The zener CR1 regulates the gate voltage of Q1 to a safe value. Capacitor C1 filters any rf or audio voltage from the signal.

The field effect transistor is connected as a voltage multiplier. It is normally

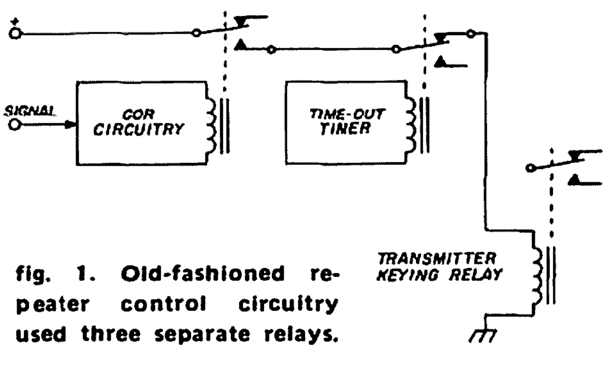
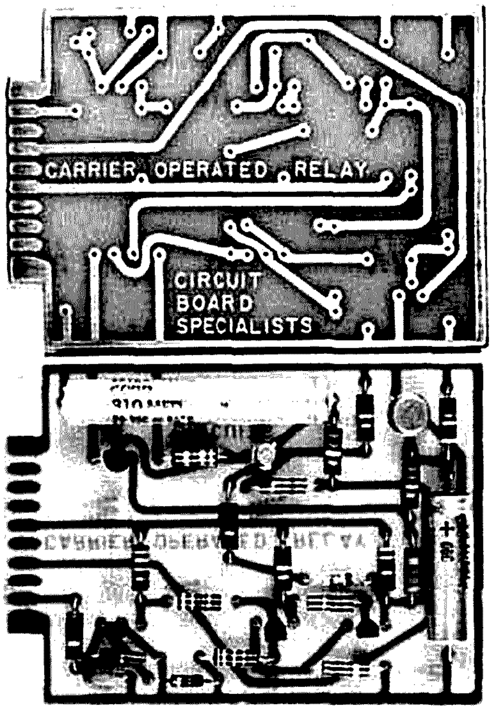


fig. 1. Old-fashioned repeater control circuitry used three separate relays.

conducting the voltage from R3 to ground causing a low voltage on the base of Q2. A negative voltage applied to the gate of Q1 will cause a pinch-off action in Q1 and consequently a rise in voltage on the base of Q2. Transistor Q2 will now start to conduct, and a rapid rise in voltage on the emitter of Q2 and a drop



The complete COR fits easily on the small pc board. Note the generous component spacing for simplicity and ease in trouble snooting. The board fits a standard 10-pin circuit-board socket.

in base voltage of Q3 will cause Q3 to cease conducting. This action of Q2 and Q3 is known as a Schmitt trigger. When the collector of Q3 goes positive as previously noted, this voltage is fed to:

1. The base of Q4 which will go positive and Q4 will conduct heavily causing the relay to close.

2. The emitter and base 2 of the unijunction transistor Q5. This starts the time-out function as will be explained a little later.

3. The anode of the silicon controlled rectifier, CR4.

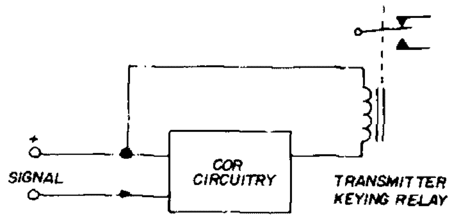


fig. 2. The updated carrier operated relay uses only one relay but offers reliable transmitter keying and built-in time-out timer.

time-out function

Now to the time-out function. As stated previously, on receipt of a signal, voltage will be applied to the emitter and base 2 of Q5. Emitter voltage, initially being very low or zero, will gradually increase, at a rate controlled by resistor R13 and capacitor C3. By juggling these values the time for the actual time-out function to take place can be set from a second to as high as several minutes. When the emitter voltage finally increases to a point where it is equal to the voltage applied at base 2, the unijunction will then conduct or fire thru base 1 and to the gate of the SCR. This gating action of the SCR will now conduct all voltage applied to the anode to ground thereby removing the base potential on Q4 and releasing the relay. The SCR will continue to conduct until its anode voltage is removed. This can only happen if the incoming signal is dropped from the receiver, at which time the COR will resume its idle state, and the voltage at the junction of R10 and R11 drops to zero thereby ungating the SCR and putting the complete system back to normal.

We have explained the action of the circuit in closing the relay and the reverse of this action will naturally open the relay with a slight delay in opening caused by the stored voltage in C2. This stored voltage prevents relay chatter on a fluttering signal and provides a tail on the repeater carrier.

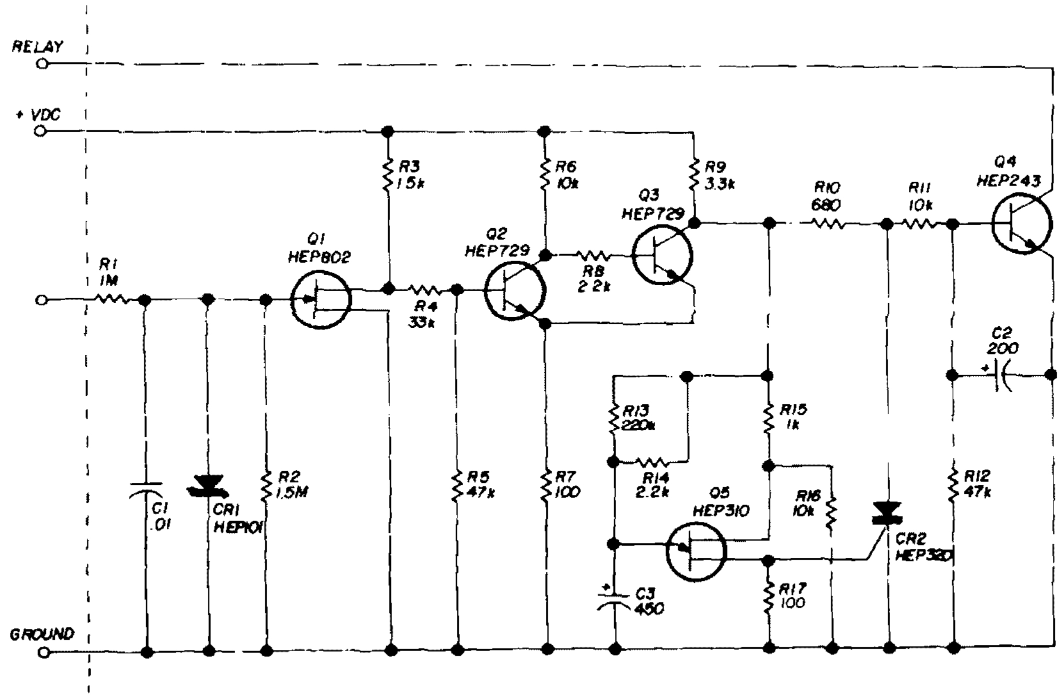
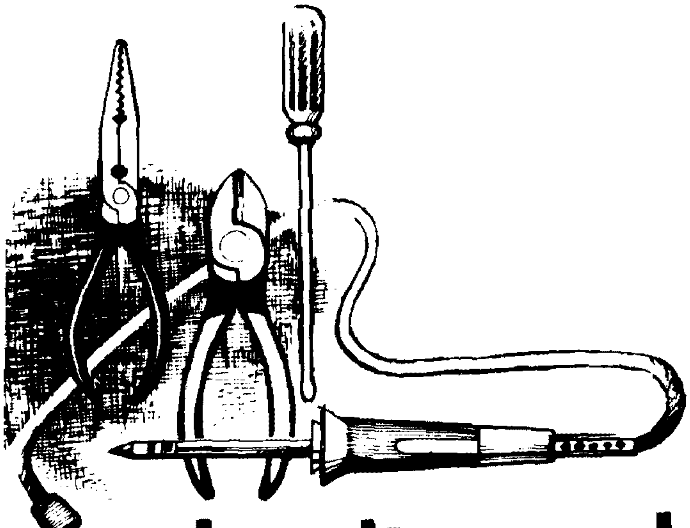


fig. 3. Schematic of the carrier operated relay.

A little added feature of this carrier operated relay is contained in CR5 and R14 which allows a slow bleed-off for the time-out function. This little jewel requires the users of the repeater to drop their carriers between each transmission for a few seconds in order to reset the timer fully. This quickly trains the repeater users to allow a little time between each transmission to allow for breaks as necessary.

ham radio



circuits and techniques

ed noll, W3FQJ

fet biasing modes

The field-effect transistor, although its operation is entirely different from that of a vacuum tube, has family characteristics quite like that of a pentode. Biasing methods are similar too.

A junction field-effect transistor consists of a channel of P- or N-type silicon semiconductor sandwiched between a region of opposite sign. Discussion is confined to an N-channel junction fet, fig. 1. In this case, the N-channel has its drain at one end and source at the opposite end, sandwiched by a piece of P-type semiconductor material. The N-type material has electron or negative carriers; the P-type positive carriers.

When there is zero bias between gate and source, the application of a positive voltage between drain and source causes the electron carriers of the channel to move between source and drain. As the positive voltage is increased this current rises. In effect, the channel acts as a resistor.

As this current rises a voltage gradient appears along the resistor and the junction between gate and channel becomes reverse biased. This causes a depletion area to extend outward from the gate into the channel. Its action is to decrease the effective cross-sectional area of the channel – similar to decreasing the diameter of a conductor or a resistor. The greater the potential gradient along the channel, the further the depletion area extends into the channel. Eventually a point is reached at which there is no significant increase in drain current because the extension of the depletion area balances out the influence of the increase in drain-source voltage. The current is then said to be pinched off.

The above explanation is represented by the Vgs = 0 curve of the family shown

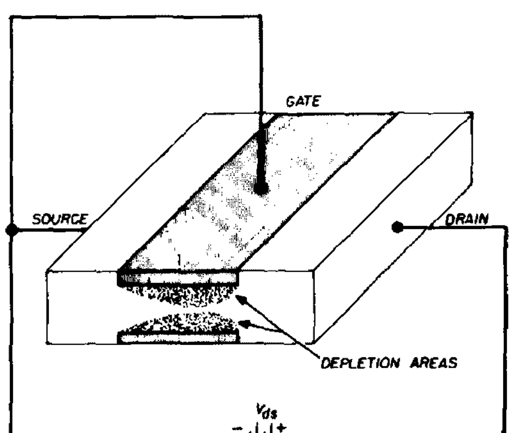


fig. 1. Simplified plan of an N-channel junction fet.

in **fig. 2**. Note how the drain current increases between drain source voltage of zero and +5 volts. The pinch-off region *occurs above a drain* voltage of 5 volts. Note how slowly the drain current now increases for a given increase in drain voltage, producing a pentode-like characteristic curve. The drain current at which pinch-off starts is known as the saturation current and is symbolized by Igss.

Note from **fig. 2** that the drain source voltage Vds corresponding to the saturation current Idss is +5 volts. The bias voltage Vgs needed to cut off the drain current (one nanoampere) for this amount of drain-source voltage is known as gate-source pinch-off voltage Vp. In **fig. 2** this value would more likely fall somewhere between −1.8 and −2.0 volts Vgs.

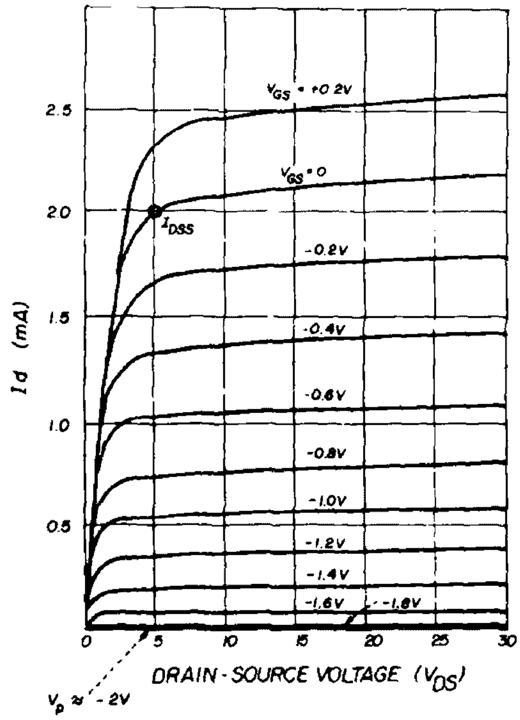


fig. 2. The family of curves for a fet.

When the gate is biased negatively relative to the source, the pinch-off occurs at a lower value of drain current because of the depletion area already set up by the gate bias. The higher the gate bias, the lower is the pinch-off drain current. A family of curves demonstrates the characteristics of a given fet as shown in **fig. 2**.

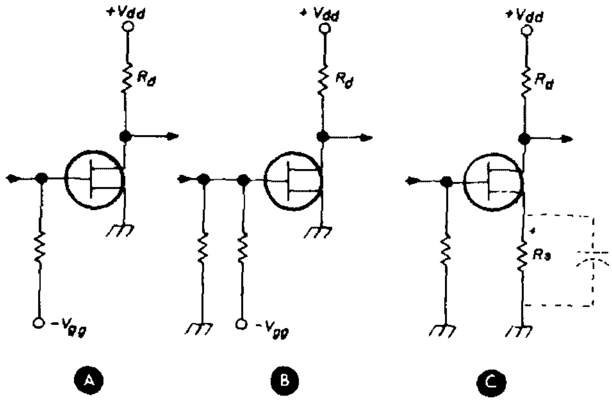


fig. 3. Simple biasing methods for fets.

simple bias methods

Three simple bias systems are given in **fig. 3**. Note their similarity to vacuum-tube practice. The first case is biased from a battery or other bias source. The gate resistor has no influence on the bias if the stage is to be operated in a normal linear fashion. In B a two-resistor voltage divider produces the correct bias. Hence it can be derived from a higher voltage source.

The third arrangement is called source bias. Drain current in the source resistor now develops a positive voltage between source and common. The current direction, as shown, is such that the gate is made negative with respect to the source. The cathode resistor of a vacuum-tube circuit performs in the same way. A capacitor connected across the source resistor prevents degeneration and loss of gain.

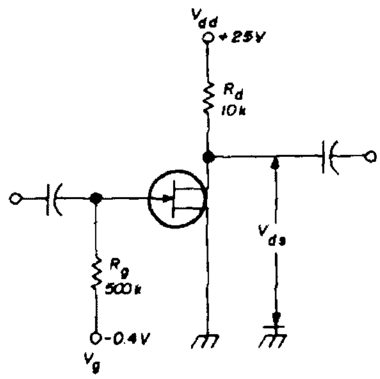


fig. 4. A circuit with external bias to meet the conditions of fig. 5.

load line

The operation of the circuit of fig. 4 can be disclosed by drawing an appropriate load line on the family of curves, fig. 5. As shown the supply voltage is +25. If on the peak of the input signal, the drain current is to rise to the saturation current on the zero-bias curve, the load line must be drawn between this point and +25 on the zero drain-current axis. If the load line is continued to the zero drain voltage axis, it is possible to determine the ohmic value of the load line:

$$Rd = \frac{25}{0.0025} = 10{,}000 \text{ ohms}$$

This load line condition is matched by connecting a 10,000 ohm resistance in the drain circuit, fig. 4.

A suitable operating point can be found along the load line. The best gain is obtained by using a low bias provided the input signal is not great enough in amplitude to overdrive the transistor. The operating point of fig. 5 is satisfactory for an input signal of 0.4 volts peak. Operating point bias is −0.4 volts; drain current 1.35 mA and drain voltage, about 11.25 volts. An operating-point gate bias of −0.4 volts is applied externally in the circuit of Fig. 4.

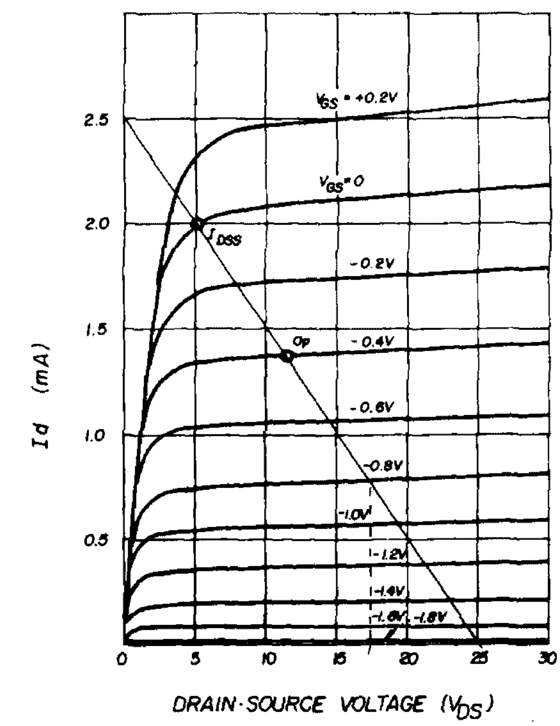


fig. 5. Fet family of curves with load line.

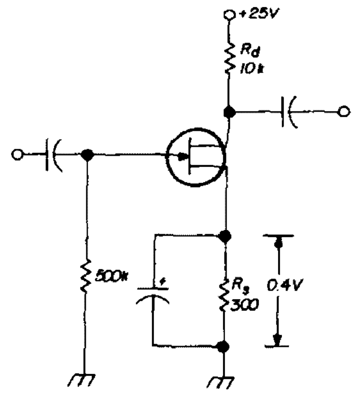


fig. 6. Self-bias circuit for the conditions of fig. 5.

If the input signal is of 0.4-volts peak, the gate voltage will swing between 0 and -0.8 volts. Hence the drain voltage swings between +5 and +17.5 volts. From this information it is possible to obtain the voltage gain of the fet stage:

$$Vg = \frac{17.5 - 5}{0.8} = 15.6$$

self-bias

Self-bias using the circuit of fig. 6 can be used to establish the same operating conditions. Inasmuch as the operating-point drain current and gate bias are known, it is possible to determine the proper value for the source resistor. Source resistor value for the circuit of fig. 6 and the operating conditions given in fig. 5 is:

$$Rs = \frac{0.4}{0.00135} = 296 \text{ ohms}$$

stabilized bias

Two factors have a significant influence on bias requirements. One is the production spread which may be such that the saturation current Idss may have a ratio of 3-to-1 or higher. In the typical example of fig. 5, it may be that the Idss value will not be 2 mA, but may have a high of 3 mA and a low of 1 mA. Temperature drift is the second factor — both warm-up operating point

drift or a shift as a result of change in ambient temperature.

The changes in operating conditions as a result of production spread and temperature are pronounced when only external gate bias is used. Self-biasing is much more self-adjusting because a shift in operating point drain current is compensated by the change in bias produced by the change of current in the source resistor. However, even better operating-point stabilization can be obtained with a further increase in the ohmic value of the source resistor. Normally this would result in an improper bias, but a combination of gate-divider bias and a higher value of sourceibias can be used to set-up the operating-point of **fig. 5**. This is done by using positive gate voltage instead of negative, however. In fact, the circuit in **fig. 7** is similar to that used in a bipolar transistor circuit but for quite a different reason as you learned.

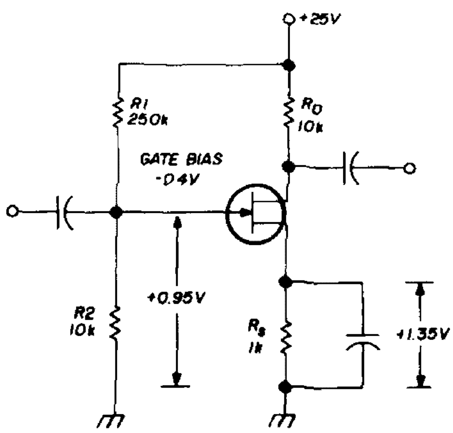


fig. 7. Compensated and stabilized bias for the conditions of fig. 5.

Good stablization is obtained by increasing the ohmic value of the source resistance three to five times. In our example, let us assume a source resistance value of 1000 ohms (an increase of +3 times). In this case the source bias would be:

$$V_s = I_d R_s = 0.00135 \times 1000 \approx 1.35 \text{ volts}$$

Note that this is substantially higher than the desired 0.4-volts bias, **fig. 5**. Hence, to obtain the proper bias, it is necessary that the gate bias be:

$$V_g = V_s - V_{gs} = 1.35 - 0.4 = 0.95$$

This condition requires that the voltage at the junction between resistors R1 and R2 be +0.95 volts.

The ratio for the two-resistor voltage divider then becomes:

$$\frac{V_{dd}}{V_g} \approx \frac{R1 + R2}{R2} \approx \frac{25}{0.95} \approx 26.3$$

A satisfactory ratio would be set up using standard value resistors of 250k and 10k, respectively, for resistors R1 and R2. This would set up a ratio of:

$$\text{Ratio} = \frac{R1 + R2}{R2} = \frac{260k}{10k} \approx 26$$

One advantage of a field-effect transistor is its high input impedance. This would be compromised by the low ohmic value of resistor R2 in certain applications. The answer to this loading of the input impedance is the circuit of **fig. 8**. In this case it is possible to use a high value gate resistor Rg. The biasing is again handled by resistors R1 and R2. However, resistor R2 will no longer load down the input impedance.

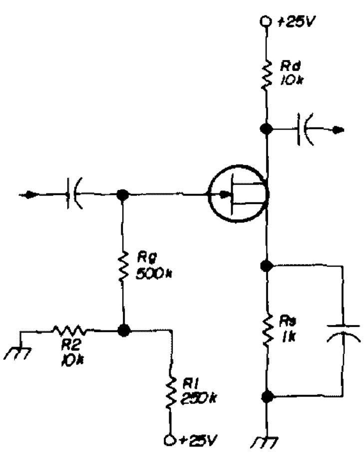


fig. 8. Circuit for the reduction of input loading.

influence of higher source bias

It is to be noted from **fig. 9** that the

actual voltage between the drain and source is less than the operating point Vds value by the amount of the source bias. In the example, this is not an important amount because of the self-adjusting characteristics of the source circuit.

In some circuits in which a rather low supply voltage, Vdd, is used and the operating point is very critical, it is advisable to increase the supply voltage by the amount of the source bias. A second alternative is to draw a load line initially that includes the ohmic value of the source resistance. Load line resistance would be Rd + Rs.

pulse-duration modulation

In a pulse-duration modulation system a low-level pulse train is modulated by the desired audio. It is then built up to a high power level by a series of switching (digital) amplifiers. They permit power amplification at high efficiency. A low-pass filter, which removes the pulse train and recovers the original modulating information, follows the final power amplifier or modulator.

This technique is being used in modern a-m broadcast transmitters with the modulator being connected directly to the cathode of the final modulated power amplifier through the low-pass filter. This plan eliminates the modulation transformer and associated critical high-power audio components. Here is a different method of modulation with which the a-m buffs can experiment. There are possibilities for the development of a kilowatt vhf-uhf a-m transmitter with less weight and fewer critical components than is common today.

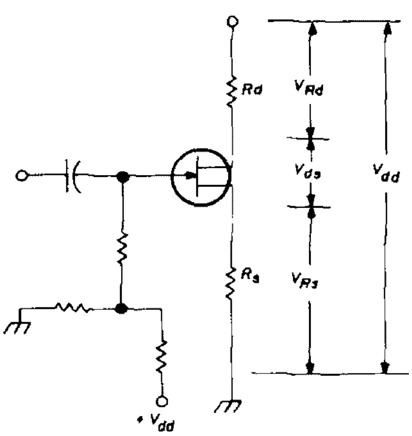


fig. 9. The figure shows the distribution of voltages in the drain circuit.

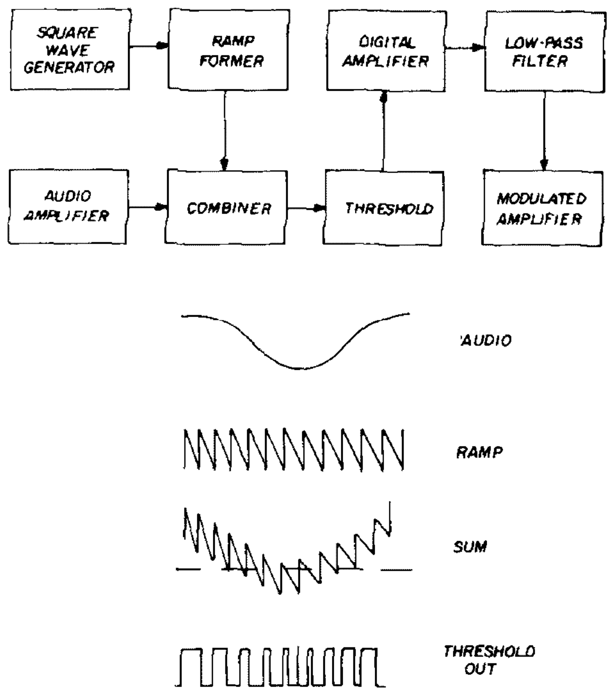


fig. 10. Block diagram and waveforms for pulse-duration modulation.

Give some thought to possible uses for pulse-duration modulation in other facets of amateur radio. Everything seems to be going digital so there should be some tie-ins. How about a digital audio amplifier and speech processor for side-band or fm?

The functional block diagram of **fig. 10** shows the process begins with a square-wave generator. This generator operates on a frequency up to five or ten times higher than the highest modulating frequency. A square wave is generated which is integrated to convert it to a ramp voltage. The waveform generator described in the last two columns could be used to generate such a sawtooth ramp voltage. The ramp voltage and the audio modulating signal are applied to a combiner stage and the waveforms of **fig. 10** show the results of this summing.

There follows a threshold stage which levels off one half of the joined waveforms. It should be noted from the waveforms that the duration of each ramp has become a function of magnitude of the original audio wave at that particular instant. Duration of the ramp for the positive peak *is* greater than during the trough of the modulating wave. Hence the output of the threshold stage becomes a train of pulses, the durations of which follow the amplitude change of the modulating wave. A change in frequency of the modulating wave simply changes the rate at which the pulse duration varies. In effect the amplitude variations of the modulating wave have been converted to a train of constant-amplitude pulses with durations that follow the original audio.

The pulse train has two levels, on and off. Hence the following amplifiers can be operated between cut-off and their full-on condition (digital switching). Efficient amplification is now possible without distorting the modulating information, and the signal can be increased in magnitude up to hundreds of watts, if desired. In a high-powered broadcast transmitter the average power level may be 20 to 50 kilowatts. In amateur practice the entire chain might be solid state, ending in an output of several hundred watts.

The final step is to integrate back to the original audio variation. This is done with a low-pass filter that removes the pulse frequency and its harmonics, leaving a copy of the original audio variation. Output can be at low impedance and will series modulate a high-powered rf power amplifier without the need for a modulation transformer and the large audio components and filters needed to supply high voltage to modulator and modulated amplifier.

TTL fm demodulator

Wireless World presented a circuit for an fm demodulator of high linearity using a TLL one-shot monostable, fig. 11.[1] Since the input of the SN74121 also includes a Schmitt trigger, no limiter is required. Muting is handled by supplying a dc component from the rectifier transistor to the inhibit inputs of the trigger. Inasmuch as the output pulse is one of constant width, the voltage developed at the output integrator is proportional to the i-f frequency. This output resistor-capacitor combination also provides de-emphasis.

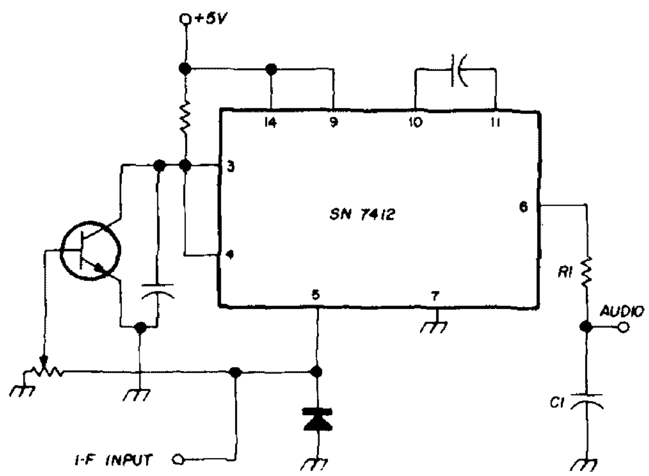


fig. 11. A TTL fm demodulator. This circuit was taken from the British electronics magazine Wireless World, April, 1972 issue.

ouch!

Several of the computer pros have clobbered some of my introductory digital IC coverage. Mr. E. D. Jensen of Honeywell sent them in good organization. I make amends by including his comments and corrections:

"In April, a typo in DeMorgan's theorem just before fig. 1 erroneously claims that X = $\overline{A \cdot B} = \overline{A + B}$.

"Figs. 1 and 2 illustrate the positive NAND function but uses the standard symbol for a positive NOR gate.

"In the June column on multivibrators, fig. 1 has several problems. Most obvious is the NOR gate RS flip-flop which is actually made of inverters. While this FF will operate correctly for the 0, 1 and 1, 0 input cases, you have wire-ORed the outputs of the FF with those of the driving gates. If these are TTL gates, the 0, 0 and 1, 1 input conditions will now destroy one or more gates.

"Furthermore, the response of a correctly wired RS FF to the 0, 0 input will always be both outputs high, not a no-change as you stated.

"The truth table of fig. 2A is also wrong in that a clocked RS FF has the same reaction to a 0, 0 input as an unclocked – at least you were consistent. The fig. 2B truth table mistakenly uses the input convention of the RS FF for the JK FF. Q is caused to go high by a 1 on J, not a 0 as is the case for S. Also, it would be much more enlightening if the table showed the outputs for 0, 0 inputs when c = 1 instead of 0.

"While fig. 5 is reasonable for an experiment, it should be noted that simply leaving TTL inputs floating is not a reliable means of insuring that they are high. Pins 1 and 2 ought to each be tied up to +5 V through 1k resistors.

"Finally, I must point out that experiment three fails to observe the 150-nsec maximum risetime spec of TTL. Many gates will indeed operate on sine wave inputs, but they are not guaranteed to do so (that is why the 7413 exists)."

errata

Count for the chains of B and C in fig. 2 of the August issue should be: 50, 10, 2 and 1 kHz and 50, 10, 5 and 1 kHz.

reference

1. Circuit Ideas, "Low Distortion F-M Demodulator," *Wireless World,* April, 1972, page 185.

ham radio

the ham notebook

noisy fans

In any equipment using vacuum tubes, the removal of excessive heat becomes imperative in obtaining reasonable equipment life. Most of the equipment presently in amateur use is vacuum-tube operated. Manufacturers of communication equipment normally *fail to remove* heat on this equipment unless the heat generated causes immediate catastrophic failure to the components (in the range of three to ten hours of operation). Yet very little is done about removing the heat which is not critical. The reason is obvious: extra parts and extra engineering increase manufacturing costs, lower profits dull the firm's competitive edge.

The solution then, is left to the consumer. Since the removal of heat is important, the best way of going about it, is by just blowing it off. The easiest way to do this is to use a small fan (like Rotron, Muffin, Sentinel, Centaur or Whisper fans). To achieve maximum efficient electronic cooling, compromises must be made. First: the amount of air to be moved should be known – the more air you move, the greater the cooling factor (assuming a stable ambient-temperature), Second: the more air you move, the more noise (air noise, rotor noise, motor noise) you generate. I found that for most amateur communication equipment, the amount of air needed for proper cooling varies from about 65 cfm to about 110 cfm. Optimum value is generally set at 75 cfm; yet it is nice to have the capability *of going to* about 100 cfm for those long operating periods or for operation on hot summer days.

Getting a Mark 4 Muffin fan and slowing the motor, I had the capability of running the fan at 75 cfm and, when things got rough, switching to 100 cfm. Establishing the proper motor speed for this type of fan, I found that the best reference setting for about 75 cfm was about 70 to 90 Vac, depending on the type of fan. You could use a speed control for the fan, but speed controls run from five to fifteen dollars. The fan only costs about ten dollars so this is an expensive method to try. If you use a power resistor in series with the fan, you are essentially generating more heat.

The best was to slow down the fan and also make it run quietly, is by using a capacitor in series with the fan. Since a capacitor does not dissipate any power (well, very little, anyway), you are thus achieving the same result.

First, determine the power consumption of your fan. It is usually indicated on a plate on the fan body. Next, determine the resistance (simplified impedance) of your fan using $Rf = (Vt)^2/P$ where Rf is the fan's resistance, Vt is the supply voltage and P is the fan's power consumption.

In my installation, I knew P (listed as 12 W) and Vt was the standard 120 Vac. The resistance worked out to be 1200 ohms.

From fig. 1you can see the voltage you want across the fan. I decided to run my fan at 90 Vac. Let this be Vf in the

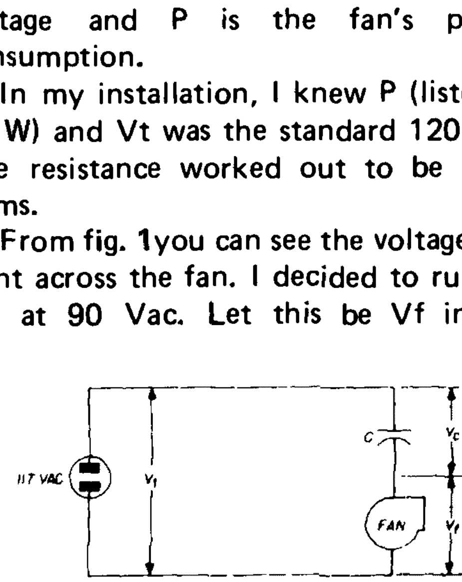


fig. 1. Voltages in the fan circuit.

figure. The total voltage, Vt, is 120 Vac. The desired voltage drop across the capacitor, Vc, is calculated simply by Vc=Vt−Vf. In my example Vc was 30 Vac.

The capacitor voltage drop, Vc, will determine the value of the capacitor needed. First calculate the necessary reactance for the capacitor: Xc=Vc/Ic. Ic is the total current through the circuit and is equal to Vf/Rf. In my case it worked out to 75 mA. By substitution, the reactance of the capacitor can be found from the new equation Xc=(VcRf)/Vf. We also know the textbook equation for reactance: $Xc=1/(2\pi fC)$.

By substitution and then solving for C, our equation becomes $C=Vf/(2\pi VcRf)$, where Vf is the voltage across the fan, Vc is the voltage across the capacitor, and Rf is the resistance of the fan at 120 Vac. My capacitor worked out to approximately 6 μF. I also placed a high-resistance carbon resistor across the capacitor to also reduce the fan speed. The switch across the capacitor allows a choice of fast or slow speed.

At 120 Vac, my Muffin Mark 4 fan moved 100 cfm with a noise level of 42 dB. At 90 Vac, the same fan moved 75 cfm with a noise level of 22 dB.

The capacitor and resistor could be mounted in a small box. Wiring is noncritical but be sure that everything is well insulated. Capacitor voltage should be 200 Vdc or better. I have tried capacitors from 1.0μF to 10μF and they all work well for *different* speeds. If you want more than two speeds, it is easy enough to switch in different values of capacitance.

Alfonso R. Torres, WB8IUF

s-line transceive mod

The purpose of patching the high frequency crystal oscillator signal from the S-line receiver to its companion transmitter is to assure accurate transceive operation. It is not practical to do this during manufacture with separate oscillators. However, an improved, fail-safe means of eliminating the crystal oscillator signal to the S-line transmitter from the receiver for transceive operation, together with a very simple calibration procedure, may be accomplished as follows:

1. Remove the high frequency oscillator patch cable from the receiver to the transmitter. The vfo cable remains in the jack marked *vfo input.*

2. Plug in the cable from the transmitter's own crystal oscillator to its *xtal osc input* as is normally done for split-frequency operation.

3. Set up the exciter with the frequency control switch in the normal transceive, *rec vfo* position.

4. Solder a 5- to 25-pF NPO ceramic or miniature air-variable capacitor across the appropriate crystal socket pins in the transmitter for the band segment to be corrected. In addition, a small value silver-mica padder may be necessary in some equipment.

Switch the receiver to *operate,* turn on the ptt switch in the transmitter, and adjust the new variable capacitor for a zero beat sync chirp which will be clearly audible everywhere on the dial. If the crystal frequency will not move in the right direction, interchange it with the one in the receiver.

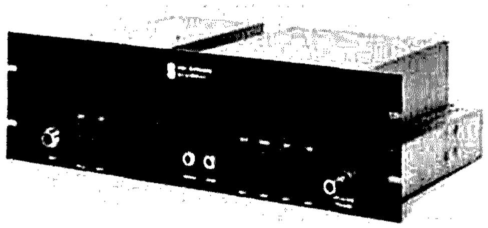
After the modification, the receiver and transmitter high-frequency fixed oscillators are corrected to approximately the same frequency so transceive operation is accomplished in a normal fashion with split operation being unaffected. The CW offset in the CW band may be essentially eliminated by the same padding procedure. Approximately 900 Hz correction is required in CW transceive to put you very near zero beat, instead of being off by the CW beat-note frequency.

In transceive operation, with this method, you now always use the transmitter crystal oscillator. This effectively precludes out-of-band ssb operation so there are no hurried patch cables to be pulled or forgotten, especially in the 14-MHz phone band when working DX below 14.2 MHz.

Marv Gonsior, W6VFR

new products

two-meter fm repeater



The new solid-state two-meter fm repeater from Standard Communications is a completely packaged repeater system including 10-watt transmitter, sensitive receiver, adjustable carrier-operated relay, time-out timer and remote control. The repeater is designed for 12-Vdc operation; current drain is 400 mA on receive and 3 amperes during transmit. Accessories include tone squelch and a 117-volt ac power supply.

The transmitter features rf output of 10 watts and frequency stability of 0.0005%. Frequency deviation is ±5 kHz, adjustable to ±15 kHz. Spurious outputs and harmonics are 55 dB below the carrier. The built-in carrier-operated relay is adjustable from noise level to 20-dB quieting sensitivity plus 10 dB. Carrier delay is adjustable from 0.1 to 10 seconds, and the time-out timer is adjustable from 0.1 to 5 minutes. An input is provided for automatic identification.

The sensitivity of the receiver is 0.5 μV minimum (20 dB quieting method). Squelch threshold sensitivity is 0.3 μV minimum. Adjacent channel selectivity at 30 kHz is 90 dB, while spurious and image rejection is -80 dB. Audio output is 5 watts; an auxiliary local speaker is available.

The new SC-RPT-1 19" rack-mounted two-meater fm repeater is priced at $640.00. For more information, write to Standard Communications Corporation, 639 North Marine Avenue, Wilmington, California 90744, or use *check-off* on page 110.

hv silicon rectifiers

A very extensive line of direct plug-in replacements for most popular mercury-vapor and vacuum glass-type high-voltage rectifiers has been introduced by Semtech. The new solid-state rectifiers are smaller and have a much longer life-expectancy than the vacuum tubes they replace. The new devices give off far less heat, require no filament power or warm-up time and are corona free. Cooling fins built into the units help conduct away heat more evenly than is possible with vacuum tubes.

Semtech calls the new rectifiers "Tubepac Silicon Rectifiers" and offers a free specifications sheet listing the PIV and maximum current ratings along with pin connection information on the complete line. The sheet facilitates direct plug-in substitution of the Tubepacs for hundreds of mercury-vapor and vacuum-tube rectifiers without any form of an adaptor.

More information is available from Semtech Corporation, 652 Mitchell Road, Newbury Park, California 91320 or by using *check-off* on page 110.

More Details? CHECK—OFF Page 110

censor-lock

The Censor-Lock is a switch which screws into the telephone handset between the carbon microphone button and the contact spring. The unit is sold to allow the telephone user to cut off audio entering a telephone line from his position *without* disconnecting the caller (as is generally done on office telephones with "hold buttons"). The unit is sold for home and office use for keeping comments, queries and confidential conversations within a room from being transmitted along the telephone lines.

The unit, however, has a particularly useful application in running phone patches by amateur radio. Often, when running a patch, there is more than enough noise and interference on frequency without having to add any more. In a normal telephone handset, noise picked up by the microphone will be heard in the speaker *along with* the caller. With the Censor-Lock in the line, stray room noises will not add to the speaker output and readability will be improved over the normal telephone noise levels.

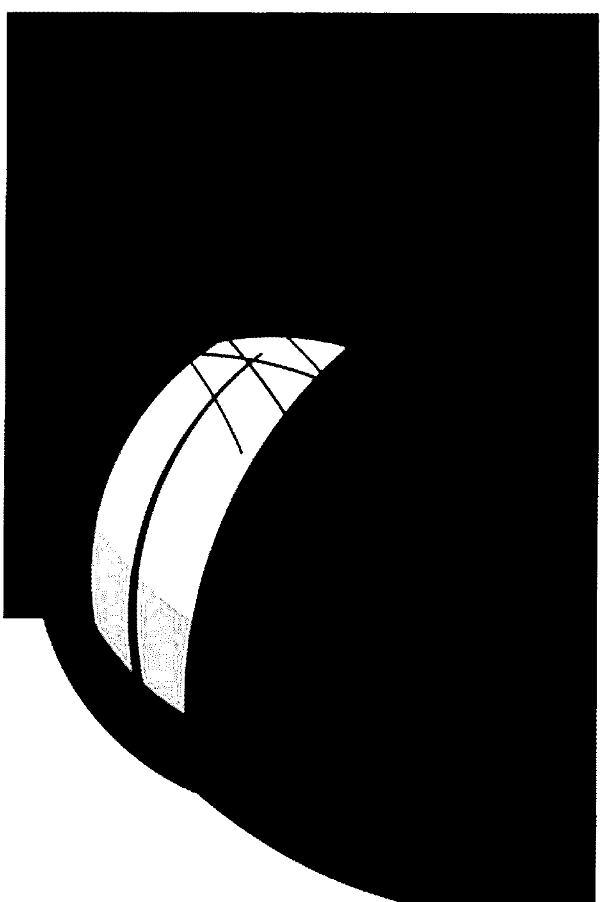
The units come in colors to match most standard telephones. They screw in without tools and can be switched in and out instantly. They sell for $4.95. More information is available from A-Head Products, Box 817, Lomita, California 90717 or from *check-off* on page 110.

75 cents

focus on communications technology . . .

ham radio magazine

DECEMBER 1972



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ham radio magazine

contents

a second look

by jim fisk

Periodically, a story about the first ham station makes the rounds. It goes something like this: In Boston, just before 1910, there were three young wireless operators, Albert S. Hyman, Robert Almy and Reginald Murray. These young men put together a small wireless station, and since there were no licensing regulations at the time, they decided to call it the Hyman-Almy-Murray Wireless Station.

They soon found that that was quite a fist-full on CW, so they took the first two letters of each name and Station HYALMU went on the air. They used this callsign for several months, but were nearly involved in an international incident when a Mexican ship named the HYALMO almost went aground off the New Jersey coast. They decided that their HYALMU callsign was too close to HYALMO for comfort, so they took the first initials of the three names, and put Station HAM on the air. The first ham station? No. But probably the first, and possibly the only amateur radio station with the HAM callsign.

Several sources have labeled this story as outright fiction, but it has been just persistent enough to arouse my curiosity. Several years ago I decided to track it down, and to determine once and for all if the story had any grain of truth.

Since at least one of the young men was supposedly a student at Harvard University and a member of the Harvard Wireless Club, I decided to start my search in the archives at Cambridge, Massachusetts. Sure enough, deep in the yellowed files there was an entry: Dr. Albert Salisbury Hyman, A.B., 1915; M.D., 1918. Could this be the same man who gave his last initial to Wireless Station HAM?

It looked promising. The Harvard Alumni Records Office revealed that Dr. Hyman was alive and well, and furnished me with his current address. I wrote to him, and his gracious reply confirmed that, yes, he was the same person who, with boyhood friends Robert Almy and Reginald Murray, had put Wireless Station HAM on the air. He also revealed that the "original ham" story got its start in a story written by wartime correspondent Percy Greenwood for a New York medical publication.

According to Dr. Hyman, Station HAM was not located at Harvard, as Greenwood's story indicates, but was actually at the Roxbury High School, which in the early 1900s was a prep school for the Ivy League.

A further search through the records disclosed that Dr. Hyman, before his graduation, was a shipboard wireless operator for the Eastern Steamship Line that ran ships from New York to Boston through the Cape Cod Canal. After graduation from medical school Dr. Hyman became a heart specialist and owned one of the first electrocardiograph machines in New York (in 1923). He was also the inventor of the artificial pacemaker used in resuscitating the dying heart.

So goes the saga of Wireless Station HAM. It was definitely not the first amateur radio station — that honor goes to some unknown wireless operator at least 20 years earlier. Nor does it have anything to do with the fact that radio amateurs are called "hams." That term goes back to the early days of wire telegraphy when unskilled, incompetent operators were called hams by their more experienced colleagues. The connotation is less than desirable.

Jim Fisk, W1DTY
editor

polarization of satellite signals

Some background for understanding polarization losses with practical antenna information for better satellite reception

Katashi Nose, KH6IJ, 4207 Huanui Street, Honolulu, Hawaii 96816

With the upsurge of interest in amateur satellite transmissions, amateurs are becoming aware of the problems and challenges of proper antenna polarization for optimum signal reception. The subject is vast and there is still a great deal of experimentation to be done. I hope that these few introductory notes and the reports on some of my experiments might be of interest and assistance to others.

Faraday rotation

Back in 1845, Michael Faraday, the experimental genius, discovered that when a block of glass is subjected to a strong magnetic field, it becomes optically active – it is able to polarize light and, conversely, is able to detect the plane of polarization of light. Light is an electro-magnetic wave and so are radio signals.

A linearly, polarized signal (from a dipole or equivalent) originating in space rotates as it travels through space, the amount of rotation being a function of the magnetic field and electron content of the intervening medium.

Fig. 1 shows a linear wave originating in some form of synchrotron radiation (radio signal) which is rotated as it progresses towards earth through space.

At the earth end of this ribbon, the receiving dipole or array must be rotated to match the polarization at that instant in order to get maximum pickup. A simplified relationship in the case of light is given by:

$$\emptyset = KHL$$

Where:

$\emptyset$ = the angle of rotation,
K = the medium through which the wave travels, sometimes called the Verdet constant,
H = the magnetic field strength and
L = the thickness of the medium.

A more practical simplified relationship in the case of radio frequencies is given by:

$$\phi = \frac{K}{f^2} NH \cos\theta \, dh$$

where:

ϕ = amount of rotation,
K = proportionality constant,
f = frequency,
H = magnetic field
θ = angle between the plane of the incident wave and the magnetic field and
dh = differential element of the path length.

It follows that you can measure the electron content of the intervening medium by measuring the Faraday rotation. Note that rotation is frequency sensitive.

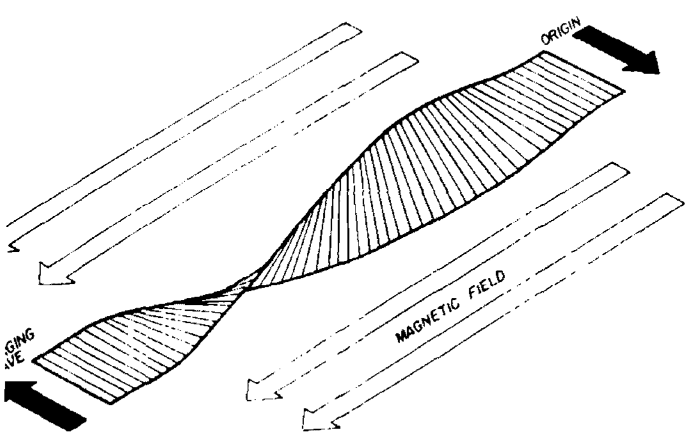


fig. 1. Rotation of wave through a magnetic field.

The photograph shows the ATS-1 *communications* satellite which weighs 800 pounds and is spin stabilized at 100 rpm, its axis of rotation being parallel to the earth. it is geosynchronous at 149° West longitude at the equator which places it around 72° elevation from Hawaii.

The top edge of the photo shows the eight vhf antennas which are phased electronically so that the array sends a conical beam with an aperature roughly corresponding to one third the earth which it illuminates.

The signal is linearly polarized, yet when it arrives on earth, it might have undergone several rotations because of the Faraday effect.

The number of rotations varies from a few degrees per hour to as much as five turns per hour and sometimes even more. A typical rotation curve is shown in **fig. 2.** Depending on the type of receiving antenna used, signal differential between planes of polarization can be as much as 15 dB. The implication is obvious: you

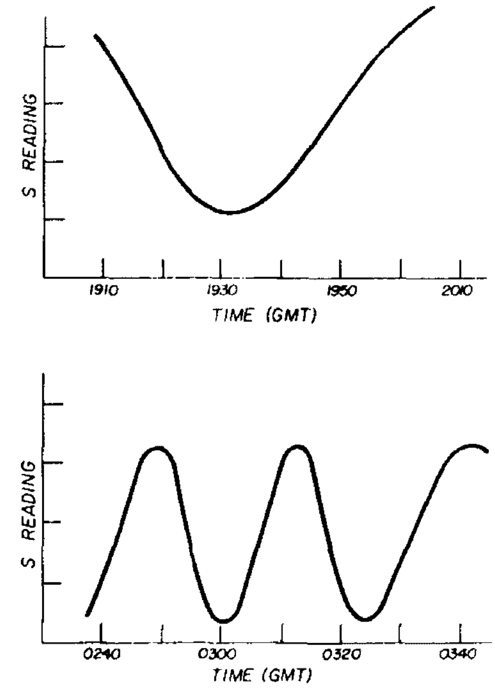


fig. 2. Signal variation due to Faraday rotation. Both *readings were taken the same day* — November 1, 1971.

must correctly polarize the antenna for successful reception.

If a piece of thin mica, commonly known as a quarter-wave plate, is inserted

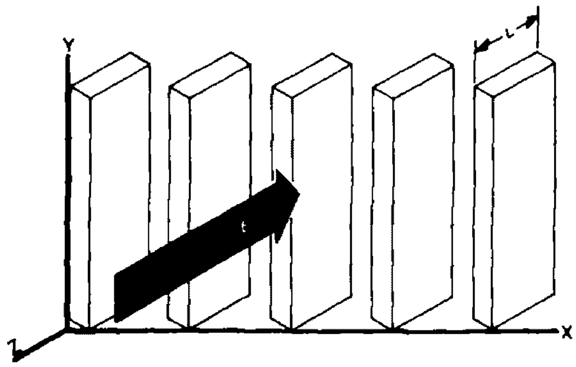


fig. 3. A wave polarizer — the E vector at 45 degrees emerges as a circular wave.

in the path of polarized light, it will introduce a 90° phase shift and a half-wave plate can cause a 180° shift.

Likewise, consider fig. 3 which shows a wave polarizer, which is merely a collection of conducting slabs. If a linearly-polarized radio wave is incident at 45° into the page (z direction), the incident electric field can be resolved into two components. The x component is unaffected since there is little interaction with the slabs in the x direction, but the y plane wave is reduced in velocity. If the L dimension of the slabs is sufficient to retard the voltage by 90°, the wave emerges from the back side (behind the page) circularly polarized. Conversely, a circularly-polarized wave originating on the back side of the page will emerge out of the page as a linearly-polarized wave.

If the slab dimension L is increased to 2L (180° phasing), then the emerging wave will again be linearly polarized since the x and y voltages will be opposite in phase, but the polarization will be displaced by 90°. Increasing the slab dimension to 3L makes the emerging wave circularly polarized again but with a counterclockwise rotation. Finally, if the slab dimension is increased to 4L, the emerging wave is linearly polarized at 45° as in the incident wave. Now, if you are confused, think what you have to contend with when fabricating a circularly polarized antenna using linear (Yagi) elements. Sometimes, at microwave frequencies, slabs of plastic are used to introduce the proper phase delay.

Incidently, the picket fence analog to explain polarization, so common in high school textbooks, is pedagogically wrong. Fig. 4 shows a rope being vibrated in a circular mode (unpolarized light) which when passing through a slit (polarizer) emerges as a polarized wave. This polarized wave passes through a parallel slit (analyzer) but not through a perpendicular slit.

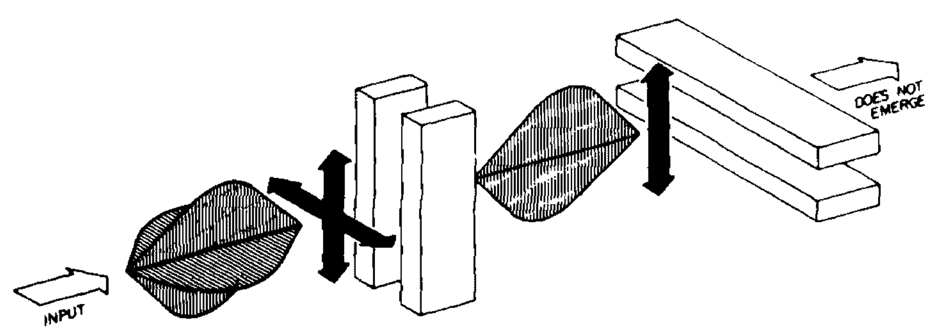


fig. 4. The erroneous picket fence analog of a polarized electromagnetic wave.

The student erroneously *thinks* of *the* plane of the opening as passing the polarized wave. Consider fig. 5 which shows an unpolarized wave from a micro-

wave generator impinging on a grid of wires held at right angles to the plane of polarization. By the picket fence analog the wave should be stopped by the second fence, but no, it goes right through without attenuation. But if the polarizer is held parallel to the plane of the wave, the wave does not emerge from the other side. If the grid elements are less than a half-wave long, there is no interaction with either orientation. Grid elements longer than a halfwave behave like a halfwave grid. The phenomenon is one of absorption and reradiation not of transmission as implied by the picket fence analog.

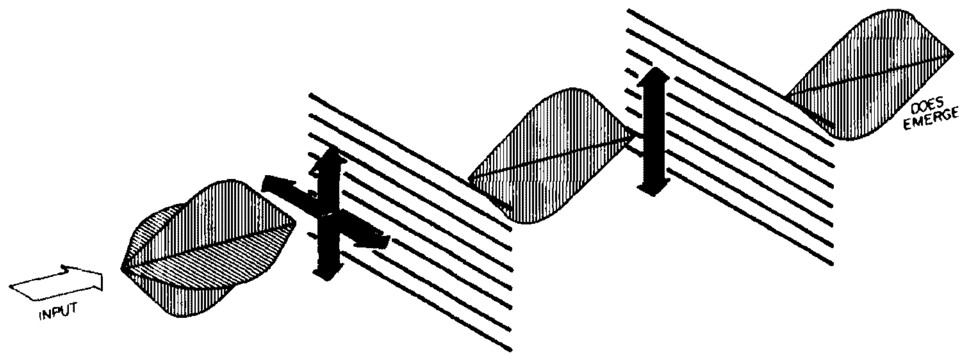


fig. 5. An electromagnetic analog of polarization.

Where does all this discussion lead to? Merely, that fabricating a circularly polarized antenna of the crossed Yagi type for picking up vertically and horizontally polarized signals is not an easy task, and that what most of us like to think of as being a circularly-polarized, crossed Yagi is wishful thinking.

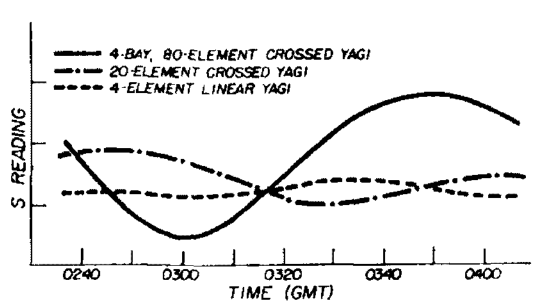


fig. 6. Comparison of different antennas during changing polarization.

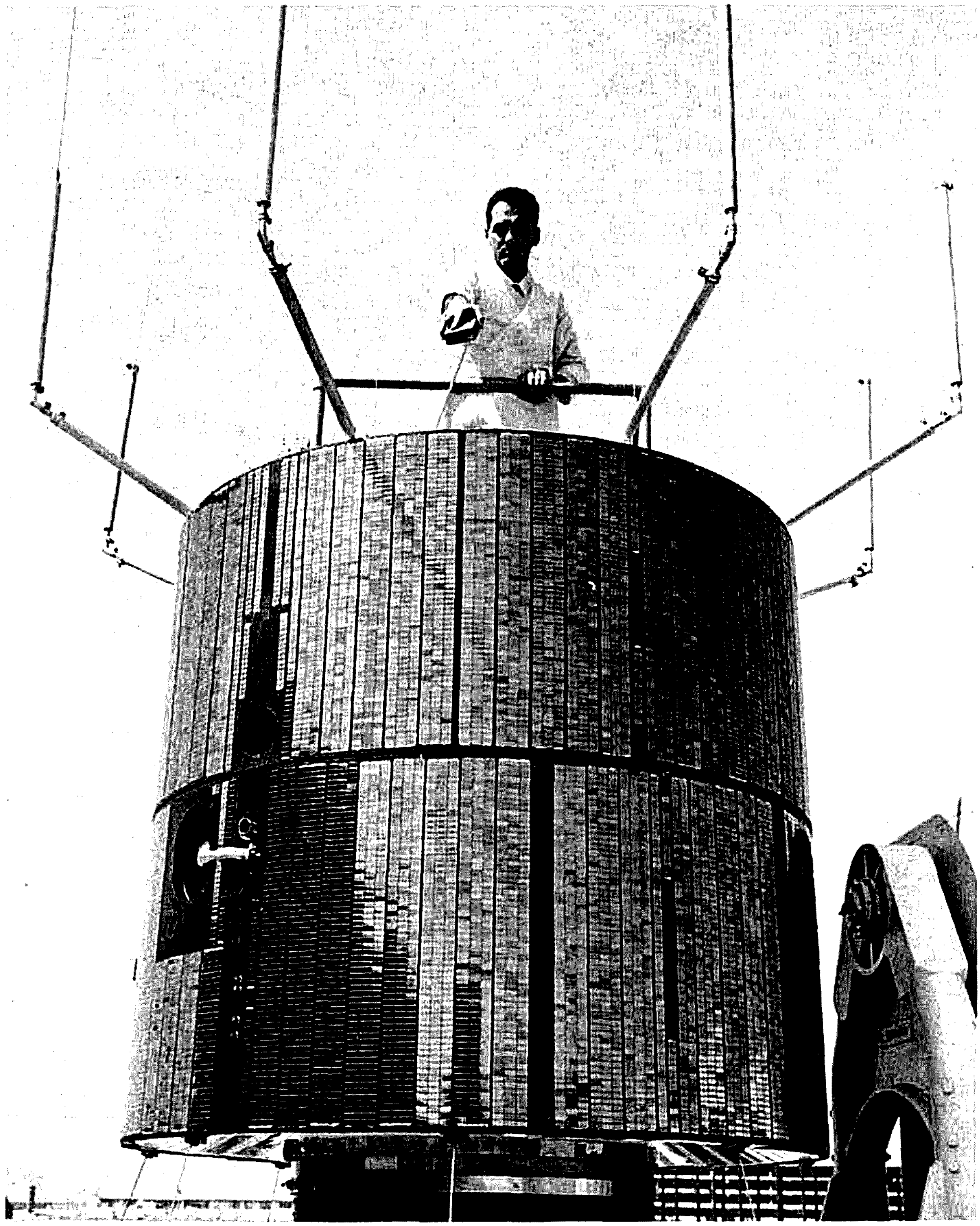
Look at what you're up against. First, each individual Yagi must produce exactly the same magnitude of voltage at its feedpoint which means that it must be electrically balanced – a difficult task at best.[1] The phasing sections must introduce the right quadrature voltage. Unless you have a line stretcher or some lumped circuit equivalent, this is difficult to do.

More basic than matching and phasing is that you lose a precious 3 dB by the mere act of phasing the two antennas. Depending on the phase and amplitude relations of the two antennas, the combination rejects some circularly polarized waves. Only one-half the energy is accepted.

practical antenna performance

Fig. 6 shows the non-circularity of a commercially made, inexpensive, 4-bay 80-element antenna designed to pick up either vertical or horizontal polarization. Also shown on the same graph is the performance of a home-made, 20-element crossed Yagi, and a four-element linear Yagi which was rotated to follow the Faraday rotation. Readings were taken in rapid sequence on the transponded signal from ATS-1.

Where does all this discussion lead to? Nothing more than what we already know – that the simplest is best. Unless you have access and the know-how to adjust a circular antenna configuration, better stick to a simple Yagi for amateur satellites. If you think big, you won't get there in time.



The Applications Technology Satellite (ATS-1) shows antennas at top. (Photo courtesy Hughes Aircraft Company.)

transmitting antenna

Unfortunately, the solution to the uplink problem is completely different. Here we have no choice but to use a circularly-polarized antenna. The entree into the satellite (assuming a linearly polarized antenna) is contingent upon correct polarity because we have no way of telling what the correct polarization is until we send up a transponding signal.

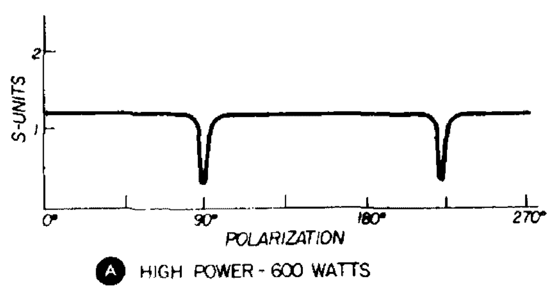
Fig. 7 shows the rotation necessary for entree into ATS-I as a function of polarization. As can be expected, with high power (one kilowatt output) there is a broad entree. With low power (ten watts), however, you have to be sure the polarity is correct.

A crossed Yagi is really a turnstile antenna shown in **fig. 8**, but arranged so that there is axial radiation by use of directors and reflectors. Note that the two radiators must be fed equal-magnitude voltage of the right quadrature – 90° lead or lag for right or left polarization. Any unbalance, either in phase or magnitude, results in an elliptical pattern which degenerates to a linear pattern in the extreme case.

Since you have no choice, make the best of a difficult situation and attempt to tune the Yagis for equal current in the driven elements. This is discussed in a previous article.[1]

The cheap and dirty method of tuning for balance is to feed considerable power into the antenna and to go around with a pencil on the end of an insulated stick and touch the various elements. By considerable juggling of element lengths as discussed in the previous article, you can get some semblance of circularity.

Meishin Electric Company of Tokyo has come up with an interesting variation. The driven elements are of turnstile configuration, but the reflectors and directors are circular discs of the proper diameter. Helix antennas are a tale in themselves.



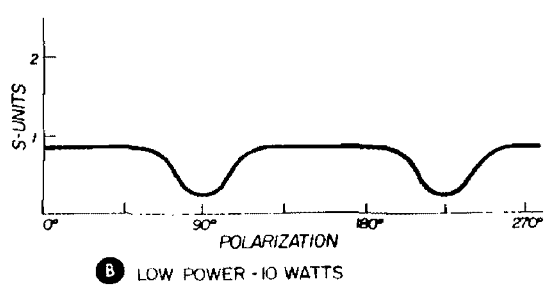


fig. 7. Results of the uplink polarization test with entree to the ATS-1 satellite. Tests were run using an 11-element linear Yagi at experimental station KB2XXK on December 13, 1971.

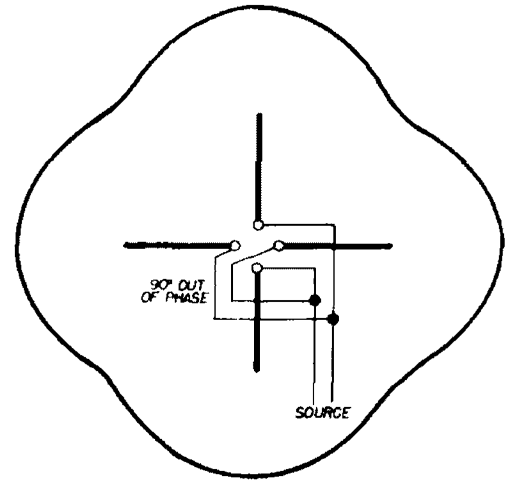


fig. 8. A turnstile antenna and the resultant pattern.

space diversity

There is little to be gained by space diversity since fading is not due to ionospheric reflection or absorption, unless you are receiving at a low elevation angle. Deep fades which occur occasionally are still unaccounted for.

acknowledgements

I wish to thank Mrs. Mary L. Burton, KH6HGO, a graduate mechanical engineer, who patiently took data for months and processed it, and to Mr. Fred Matsunaga, PhD candidate in physics, who stuck with us through the first few months when we did not know which way we were flying.

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ham radio

high-performance single sideband rf speech clipper

This solid-state rf speech clipper features the use of crystal filters to provide maximum talk power with minimum distortion

Leslie Moxon, G6XN, Hampshire, England

As I pointed out in last month's issue of *ham radio,* there is considerable confusion over the various aspects of speech processing for ssb.[1] I discussed the different processing systems that are currently being used by amateur radio operators, and tried to clear a way through some of the conflicting statements that have permeated the literature. This month I will discuss a practical ssb speech-processing system, along with some circuit details and operating advantages.

practical speech processing

The first systems I tried were based on surplus FT243 crystals built into a half-lattice filter as shown in fig. 1. Some crystal selection was necessary, particularly for the shunt crystals (Y1 and Y3) which produce the notch on the carrier side of the sideband. Inductor L1 was wound by cut and try to resonate at the desired i-f with about 200 pF at C1. The secondary of L1 consists of 5 or 6 turns wound symmetrically over the primary.

The value of resistor R2 is fairly critical, and if you have trouble obtaining proper passband response, this is the first component to change. If you are not able to obtain suitably-spaced commercial crystals, you may have to resort to grinding. However, the 2.2 to 2.5 kHz frequency separation required is fairly small, and crystals should be readily available.

Although I used an i-f of 5435 kHz, the same alignment procedure should be satisfactory for other frequencies. I have used these half-lattice crystal filters with 2N706 input and output circuits in several transceivers, but construction is simplified through the use of integrated circuits. I used ICs for an improved rf clipper design based on 5200-kHz Japanese crystal filters as shown in fig. 2. There should be no difficulty in adapting this circuit to FT243 crystal filters, although there would be some performance loss because of the lower number of crystals.

design

The first clipping stage is operated with a low value of bias to provide a large dynamic range (and relatively poor limiting). The second stage operates with a higher bias level which provides the over-all limiter characteristic plotted in fig. 3. The diodes I used in the circuit of fig. 2 were salvaged from surplus computer boards, but any switching types should be suitable (I used 1N914s in earlier designs).

The value of capacitor C1 (about 3 pF) may need to be changed to insure that the characteristics of the limiter circuits coincide. The value of C2, about

2 pF, is selected so that U2 is not driven too hard. With a six-volt power supply, the output voltage was 380 mV, varying with the supply voltage as indicated in **fig. 3.**

solid-state ssb exciter

The rf clipper is part of the solid-state ssb exciter shown in block form in **fig. 4.** This transceiver is based primarily on Plessey SL600 series integrated circuits, and can be detached from the main rig for portable use. When running 400 watts PEP output from the main rig, thorough shielding of the solid-state exciter, and filtering of all power-supply, microphone and audio-output leads is essential to prevent rf feedback.

A calibrated gain control, R1, was included in the circuit for clipping level adjustments. This is quite straightforward since attenuation varies linearly with the voltage applied to pin 8 of the SL630 microphone amplifier, with zero attenuation at 0.8 volt, and 30-dB attenuation at 1.34 volts. The gain of the pre-amplifier was adjusted, through choice of components, to suit the sensitivity of the microphone. When completed, it was just possible to reach the maximum allowable input of 200 mV to the balanced modulator when speaking loudly.

Some pre-emphasis is provided by C1 and the 500-ohm input resistance of the SL201 speech preamp. The rf gain between the balanced modulator and clipper is preset by R2 so that normal voice peaks just reach limiting level with R1 set for 25 to 30 dB of attenuation.

A monitor scope on the output of the transmitter was used to observe limiting level. The monitor scope was also used in conjunction with the rf gain control, R3, to check the transmitter for linearity, an operation which is made easier by the relatively constant output level provided by the clipper.

It is not always convenient (or desirable) to place the rf gain control immediately after the second filter, and it is important to make sure that the intervening amplifier or mixer stage is linear. You might think that any non-linearity between the balanced modulator and clipper would simply leave the clipper with less work to do, but unfortunately, capacitors in bias and decoupling circuits have a habit of charging up so that stage gain varies with the peak or mean signal level. Therefore, it is advisable to check linearity and make sure there are no amplifier time constants exceeding about 1 millisecond.

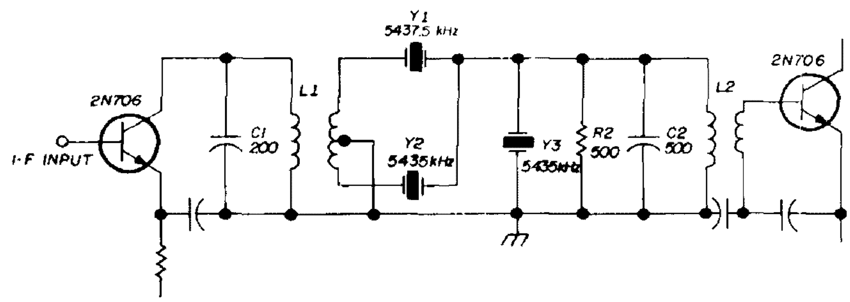


fig. 1. Half-lattice crystal filter using surplus FT243 crystals. L1-C1 and L2-C2 resonate at 5436 kHz. Crystals Y1 and Y3 are selected, as discussed in text.

performance

In use, this system provides a big increase in intelligibility when signals are weak or interference is bad. The operator's voice is always clearly recognizable by those familiar with it, and removal of all clipping produces an average drop of about two S-units. In general, other operators have no idea that speech clipping is in use, and are often quite surprised when they are told. However, the use of more than 20 dB of clipping tends to result in adverse reports.

surplus crystals

The carrier and unwanted sideband rejection of surplus crystal filters, although much inferior to commercial crystal filters, should be adequate for at

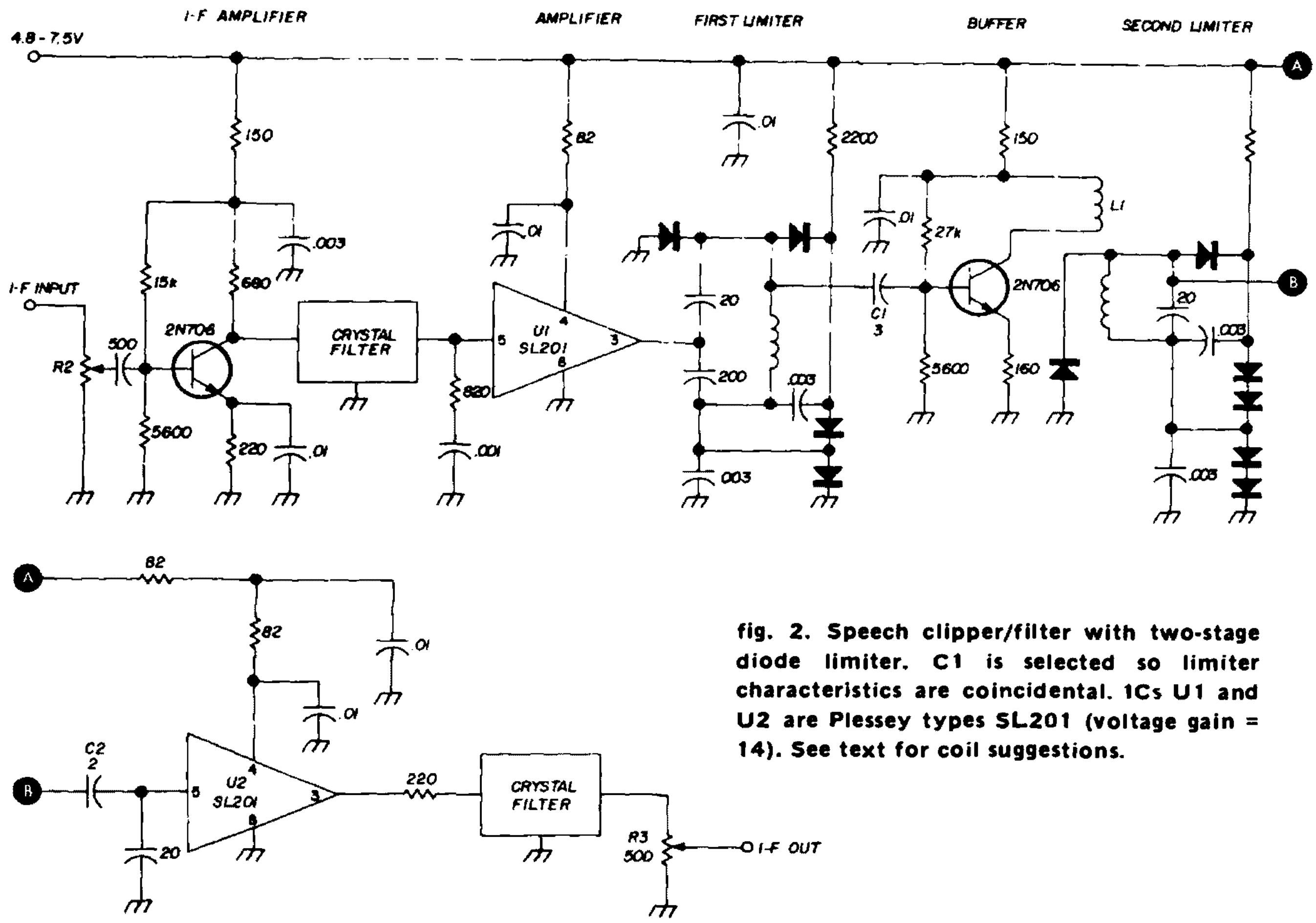


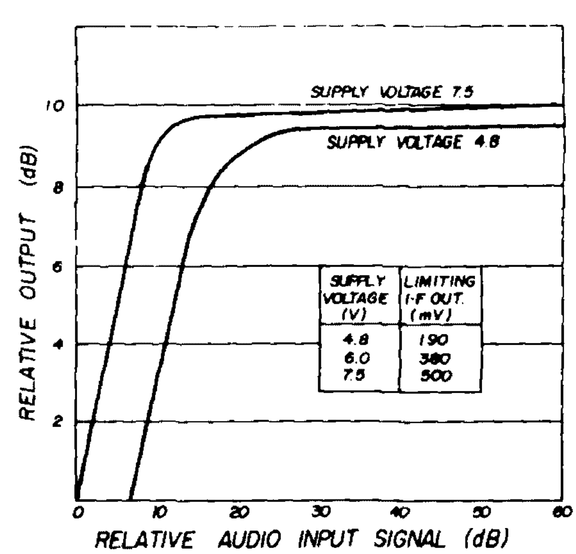
fig. 2. Speech clipper/filter with two-stage diode limiter. C1 is selected so limiter characteristics are coincidental. ICs U1 and U2 are Plessey types SL201 (voltage gain = 14). See text for coil suggestions.

least 12 dB of rf clipping. Surplus crystal filters are not normally suitable for receiving because of the high amplitude of spurious responses. However, the use of two filters for the rf speech-clipping system offered a key to the reception problem, with two separate filters. In the receiving filters the crystals were selected to stagger the spurious responses so that any unwanted frequency passing through one filter were blocked by the other.

alignment

The system shown in fig. 2 is easily checked for correct operation. A sensitive rf indicator such as that shown in fig. 4 is connected to the output of the balanced modulator. An audio tone is applied to the microphone input and the audio gain is increased until the rf signal output starts to limit. The audio gain is increased by about 20 dB (with R1) and the transmitter output is checked to make sure the output remains constant while the output of the balanced modulator rises linearly.

Connect the rf indicator to the output of the second filter and increase the rf at this point by increasing the size of capacitor C2. The output of the final should increase proportionately, indi-



SUPPLY VOLTAGE (V)	LIMITING I-F OUT (mV)
4.8	190
6.0	380
7.5	500

fig. 3. Limiter characteristics of the circuit of fig. 2. Curves represent extreme values of battery voltage for portable operation.

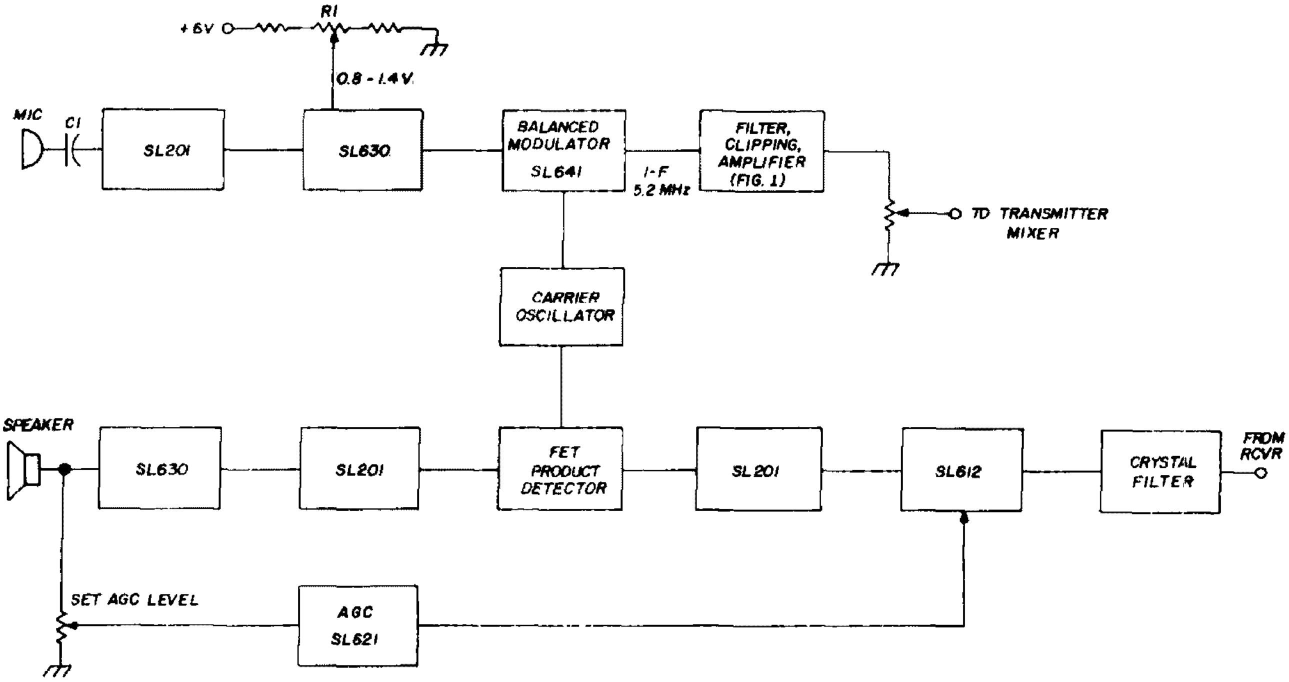


fig. 4. Solid-state ssb transceiver used by author is based on Plessey SL600 series ICs.

cating that there is no non-linearity occuring after the limiter. Alternately, if the rf gain control, R3, is not at maximum, the rf indicator may be placed after R3, and R3 varied instead of C2.

Remove modulation, place an rf short between the input and output of the second filter and check that carrier leakage remains at least 20 dB down, relative to the limiting level. Place an rf short across the input to the second filter and check that there is no rf output; repeat this test with a short across the input of the first filter.

conclusion

For the present, rf clipping is likely to appeal primarily to the serious experimenter who is looking for the ultimate in performance. There are no shortcuts to good performance, and as I noted in last month's article, careful attention to a large number of small details is essential. However, even with simple designs, you can expect some increase in talk power.

In this context, of course, I am talking about the end product, and not the means of achieving it. The system of **fig. 2** is not the only possible method. For example, by converting the clipped i-f signal back to audio, you have a device which can be plugged into the microphone lead (such as the Comdel speech processor).

Yet another approach is the modification of commercial gear along the lines used by K6JYO.[2] Each case presents a different problem, but some involve less difficulty than others. Although it is probably not always possible to follow all of the rules mentioned in my first article, particularly in regard to filters, the inclusion of as many points as possible is likely to prove worthwhile.

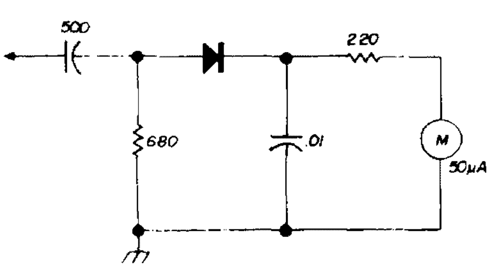


fig. 5. Sensitive rf probe for aligning rf speech clipper.

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ham radio

evaluation of simple dx antennas for 40 and 80 meters

DX is possible on 40 and 80 meters with a variety of antennas designed around practical limitations

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The introduction of awards by several organizations for recognition of multiple band DX capability have considerably diversified the antenna farm of the serious DX man. Working DX on 10, 15 and 20 meters is no problem using the popular three-band beam and an average station. Things get more difficult, though, as you move inland from the east or west coast. Due to the deterioration of signals as they are propagated over land at low frequencies, amateurs in the interior require antennas more elaborate than a dipole at 25 feet to compete with their colleagues in more advantageous geographical locations.

a proposed solution

A survey of available commercial low-frequency DX antennas revealed that most were an electrical compromise and were also expensive. There was also the problem of compatibility with the present tower-rotator-tribander configuration. I looked to see what could be homebrewed. The results were somewhat discouraging – I didn't have a tuner, a 100-foot tower, 20,000 feet of wire or a big lot. The only simple low-frequency antennas which were useful as DX antennas and could be considered constructionally feasible were the vertical, the horizontal wire beam and the sloped dipole. I developed the following test

plan: First, to construct simple wirebeam, ground-plane and sloped-dipole antennas for 40 meters, and to determine which antenna was the most effective for DX. Second, to extend the 40-meter ground plane to about 60 feet to act as quarter-wave radiator on 80 meters and to attempt to use the same antenna as something close to a 5/8-wave vertical on 40. Next, I wanted to determine how well the ground plane worked on 80 and if the 5/8-wave vertical is more effective on 40 than the quarter-wave ground plane. Last, I wanted to construct a sloped-dipole for 80 meters to compare with the ground plane for the same band. I did not try a wire-beam configuration on 80 because of *its large size and* generally unfavorable comments in the literature.[1]

two element wire beam

The *ARRL Antenna Book* contains a simple two-element wire beam in the chapter on 14-, 21- and 28-MHz antennas.[2] Extending the concept to 40 meters was easy. The original design for the two-element folded dipole beam specified that the radiating elements should be constructed of number-12 wire, spaced 3 to 6 inches. Constructing something of this nature appeared to be a lot of work, so I used regular 300-ohm twin-lead instead. As an effective flat-top configuration required four poles of at least 60 feet, I tried an inverted-vee array using a boom for proper spacing.

The boom was assembled from an old tv mast and two sections of conduit. I installed pulleys on the boom so that the folded dipoles could be pulled out to the proper *spacing from* the tower. I attached the boom to the tower as follows: I mounted an eyebolt through the center of gravity of the boom and attached an 18 inch threaded rod to the tower at the 55-foot level. The rod was attached with U-bolts, with about 6 inches to the threaded rod extending from the tower. The eyebolt on the boom was then slipped over the threaded rod and secured with two bolts.

Stability of the boom was improved by installation of a wooden beam as shown in fig. 1. The length of the wooden beam is dependent on the desired orientation of the boom. The ends of the boom drooped, of course, so I added two eyebolts five feet from either end. Cables attached to the eyebolts were connected to the tower at the 70-foot level. Axial tension was increased by turnbuckles. Once the folded dipoles were in place on the boom and the ends attached to trees, each leg of the dipoles made a 30° angle with the horizontal.

Electrically, the antenna consists of

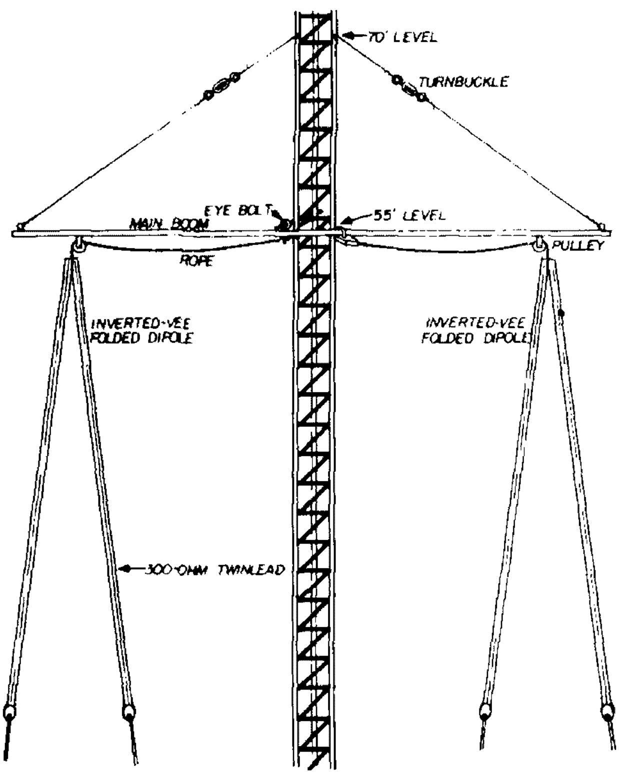
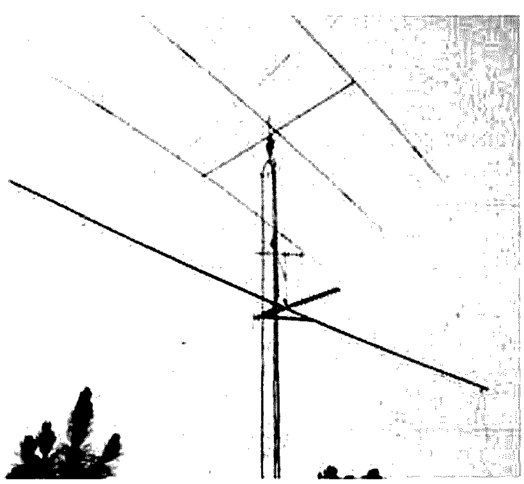


fig. 1. Wire beam for 40 meters uses two inverted-vee folded dipoles made from 300-ohm twinlead. Boom is attached to tower as shown in fig. 2.

two inverted-vee folded dipoles spaced a quarter-wave apart. The length of the feedline to the folded dipole which is the back element of the beam is a quarter-wave longer than the *feedline going to the* folded dipole which is the front element of the beam. The major radiation lobe is off of the front element. The feed impedance for the original flat-top configuration was 150 ohms. Since I didn't have any 150-ohm twin lead, and since I had no idea what the impedance would be for the modified array, I tried feeding

the antenna with 72-ohm twinlead. The swr was very high.

I tried 300-ohm twinlead and the swr was 1.5:1 at the resonant frequency. The dipoles were originally cut for 7150 kHz but the resonant frequency was about 6950 kHz. This was anticipated because inverted vees have lower resonant frequencies than dipoles for the same antenna length. Six inches removed from each end of each dipole moved the resonant frequency to 7050 kHz. The swr was 1.5:1 at 7050 kHz and 3:1 at 7300 kHz, which was acceptable. I speculated



Installation of the folded-dipole inverted-vee antenna at W5RUB. The two folded dipoles are suspended from a boom mounted on the side of the tower at the 55-foot level (see fig. 1).

that possibly the swr measured at the end of the 300-ohm transmission line and the feed point of the antenna were different, but an swr measurement at the feedpoint indicated that they were practically the same. A dpdt relay was installed at the feedpoint so I could reverse the direction of the major radiation lobe.

40 meter ground plane

Next, I built the ground plane. The antenna was made by topping off a 50-foot push-up tv mast with a 21-foot CB whip. The radiating element for 7 MHz is only 33 feet, but the additional height was necessary so the antenna could be used as an 80-meter ground plane in future tests. The mast was guyed only at the ten-foot level during the 7-MHz tests. Unfortunately, the top section of the tv mast and the bottom section of the CB whip were the same diameter so they had to be joined with a section on one-inch inner diameter copper tubing.

Ground losses for vertical radiators can be significantly reduced by using artificial grounds like a ground plane. It is wise to erect an array of this nature as far above ground as possible. One common location for a ground-plane antenna is on the peak of the roof. Many people feel, however, that radials spread out on the roof detract from the house's appearance.

To avoid this problem the 40-meter quarter-wave radials were apprehensively installed on the ceiling of the attic. Not knowing the electromagnetic properties of my roof (asbestos shingles, tar paper and plywood), I anticipated assorted gremlins, but none have been observed. Ten 33-foot radials were attached to the ceiling of the attic spaced 30 to 40 degrees apart — some being bent to fit the available area of 30 x 70 feet. If an open area is available for erection of the ground plane, by all means use it, but if yard space or aesthetic arguments are a problem, the attic provides a good alternative.

Insulating the vertical radiator from the roof was another problem. After some experimentation, the best solution appeared to be a roof saddle with a U-flange and some method of insulating the U-flange from the vertical. The best solution was an ordinary automobile rear shock absorber with rubber inserts on the mounting eyes. Using the shock absorber, a quarter-inch bolt was inserted successively through one side of the U-flange, the lower rubber insert and the other side of the U-flange. A hole slightly larger than a quarter inch was then drilled in both sides of the bottom section of the 50-foot tv mast about six inches from the bottom end (see fig. 3). The bottom section was then slipped over the shock absorber and a bolt inserted through one of the previously drilled holes in the mast, through the upper rubber insert of the shock absorber and out the other side of the mast. Thus, once erected, the

vertical was insulated from the roof.

If your roof is guaranteed to be a good insulator this ritual is not necessary, and the vertical can simply be attached to the U-flange. A hole was drilled, under a shingle, through the roof and a number-ten wire was inserted to connect the vertical element with the transmission line. Plenty of plastic roof cement was applied to the modified part of the roof. The ground plane was fed with 52-ohm coax (which was also in the attic). No reflected power was observed at the calculated resonant frequency.

the sloped dipole

The sloped dipole configuration consisted of a simple folded dipole attached between one point 47 feet up the tower and another point 47 feet from the base of the tower. Hence, it made an angle of 45° with the horizontal with the major radiation lobe for the antenna in the direction of the slope.

performance comparison

The relative merits of antennas are usually evaluated in terms of field strengths at various angles and distances from the point source of the radiation. However, the major object of this exercise was to work DX with a minimum amount of blood-letting, hence, whichever antenna had the best punch was obviously the best antenna.

The sloped dipole and one major lobe of the beam were oriented towards Europe which is the major source of long-haul DX from Mississippi. Averaged reports received from Europe indicate that the sloped dipole and the ground plane are equally effective. The average report for the two antennas ran about one S-unit higher than the faithful inverted vee at 65 feet. The two-element wire beam generated average reports which were two S-units higher than the inverted vee and one S-unit higher than the sloped dipole or ground plane. Included angles of other than 45° might be tried with the sloped dipole. No optimum angle has been published for 40 meters as far as I know.

The sloped dipole is economical and easy to erect if a 50-foot support is available. This antenna provides some reduction of stateside interference and a worthwhile increase of signal strength on distant propagation paths as compared with an inverted vee or dipole at a comparable height. The ground plane requires very little erection area (if the radials are in the attic) and provides a tremendous reduction in stateside interference. This antenna is a good omnidirectional performer for the DX man who does not have much space or has no

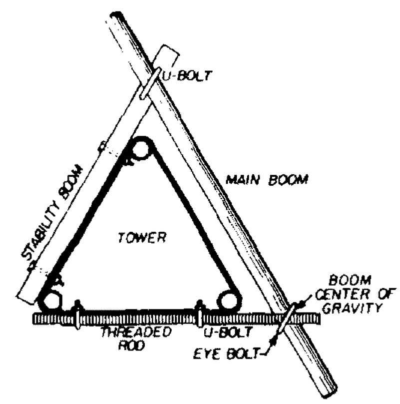


fig. 2. Boom for wire beam is attached to tower with an eyebolt and threaded rod. Short piece of wood is used for stability boom. See text for complete description.

means of raising an antenna to 50 or 60 feet.

The two-element wire beam takes up a lot of room and takes quite a bit of erection effort, but this is a good antenna to have in a pileup. Two of these antennas mounted perpendicularly to each other will provide excellent worldwide coverage on 40 meters. Rejection of signals off the side is good and fair off the back. Stateside interference, of course, increases off the front and this is sometimes a problem if the East Coast is between you and Europe. Many times I have worked Europeans who could only hear me on the wire beam, but I could hear them only on the ground plane.

On the other hand, the beam is a great contest antenna. I have often found as a stateside contest winds down on the

higher frequencies and you are forced to go to the lower frequencies in search of contacts, it is sometimes difficult to work the East Coast until a couple of hours after sunset. This is because much activity is concentrated in the 1-2-3-8 and upper 4 call areas and the signals are so strong within this area that signals from the outside are buried in the interference. The beam is a very effective aid for getting the attention of the East Coast community both in the hours near sunset and later in the evening.

dual-band antenna

As previously mentioned, the 40-meter ground plane was constructed so it could be converted to an 80-meter ground plane and something close to a 5/8-wave vertical for 40 simply by extending the push-up tv mast. The 5/8-wave on 40 is significant because at that length maximum low-angle radiation occurs.[3] The vertical was raised to 60 feet and the antenna was guyed at 10 and 28 feet. The guys were attached to eyebolts in the roof some 15 feet from the base of the antenna. The eyebolts were anchored in the rafters and plastic roof cement applied around the eyebolt on the roof surface. The vertical was insulated from the guy wires by egg insulators which were placed in each guy wire about six inches from the vertical. Six feet of the fourth section and five feet of the fifth section of the tv mast and the 21-foot CB whip were left unguyed. The 32 feet of unguyed antenna have presented no stability problems and the antenna has sustained gusts to 70 mph without damage. The lower three sections of the tv mast were extended to the maximum attainable length which came to about 28 feet total, but the fourth and fifth sections were extended to only six and five feet respectively in order to keep down the number of guy wires.

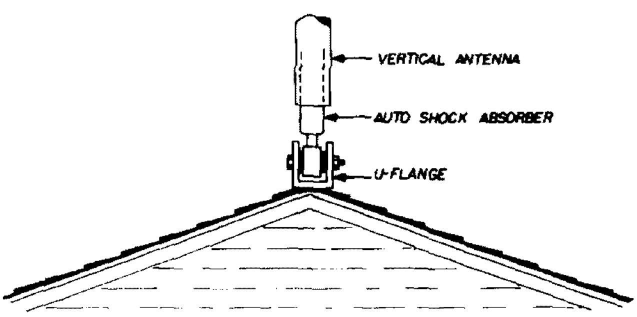


fig. 3. Vertical antenna is insulated from roof with an old automobile shock absorber with a rubber bushing.

A dpdt relay was installed at the base of the ground plane (in the attic) to switch between the 40- and 80-meter matching circuits. The 40-meter configuration required a simple L-network to match the 5/8-wave vertical to the 52-ohm coax.[4] On 80 a small loading coil was necessary because the vertical was electrically short at 3.5 MHz.

Operation of the extended vertical on 40 *indicated that no* increased effectiveness was apparent as compared with the quarter-wave vertical. This conclusion was based on comparison with the 40-meter sloped dipole. Therefore, if you don't plan to use the vertical as a ground plane on 80, the extended vertical is not worth the effort.

ground system

Contrary to popular opinion, radials shorter than one-quarterwave can be effectively used in a ground plane antenna configuration provided the vertical radiator is *at least* 1/5 wavelength or longer. No problem was experienced in loading the 80-meter vertical against the 40-meter ground system using the L-network. I worked a lot of DX on 80 using this configuration. Some weeks later, fifteen quarter-wave radials were added to the system making it a true 80-meter ground plane. This made a total of fifteen 80-meter radials and ten 40-meter radials. Reports received from DX stations were somewhat better, but I am not sure whether the improvement was due to increasing the number of radials or adding longer radials. It is interesting to note in Lee's work[3] that increasing the length of radials from 1/8 to 1/4 wavelength increases the unattenuated field for a vertical radiator at one mile by 5 *millivolts* per mile while doubling the number of

quarterwave radials from 15 to 30 provides a 12 *millivolt* per *mile* increase. Agreed, we are comparing apples and oranges, but it is conceivable that an amateur who is cramped for space may do nearly as well with thirty 1/8-wavelength radials as an amateur with fifteen quarterwave radials.

Quarterwave *radials* on 80 *take* up a lot of room. My attic, only 30 by 70 feet, necessitated placing the 80-meter radials on top of the roof. The 80-meter radial system was connected to the transmission line and the 40-meter ground system by an additional number ten wire through the roof. Fortunately, during DX season on 80 the nights are long and the days short; so if you are clever with the deployment of radials, evening guests will never be aware of the conglomeration of wire hanging above their heads. My gracious wife allows me to lay out my radials anytime after we go off of daylight savings time!

80-meter sloped dipole

Dalton recommends using a 100-foot tower to support a sloping dipole cut for 3650 kHz. This configuration yields an included angle of 52° between the antenna and ground, which he says is optimum for DX. Unfortunately, my tower is only 70 feet, and by using the same scheme, the included angle would decrease to 33°, obviously unacceptable. Practicality dictated a compromise. I decreased the included angle from 52° to 45° and raised the resonant frequency to 3800 kHz. The length of the antenna now was 123 feet. I stretched 100 feet of it from the ground to the top of the tower (at a 45 degree angle) and dangled the remaining 23 feet down the side of the tower (secured at the 47-foot level). This array worked fine at 3800 kHz but the swr was very high 100 kHz from the resonant frequency. It is well known that the bandwidth characteristics of a folded dipole can be improved by placing shorting straps at a distance from the center of the dipole which is equal to the velocity factor of the twin lead times half the length of the dipole.[5] This worked out to be about 7.6 feet from the ends of the antenna on 3800 kHz. Using this configuration, the swr was 2.5 to 1 at the low edge of the band.

80-meter antenna comparison

Reports received from Europe and Oceania indicated that the ground plane has an edge over the sloped dipole although I noticed no difference on the receiving end. The sloping dipole may be equal to, or more effective than, the ground plane if it were erected correctly utilizing the 100-foot tower.

recommendations

If supporting structures of 100 feet are not available, try the ground plane for an effective DX antenna. If a 100-foot structure is available, try a sloping dipole first, since the erection effort is small compared with the effort expended in putting up a ground-plane radiator with its associated radial system.

Possibly neither configuration is feasible; if this is the case, the modified sloping dipole is preferred over an inverted vee at 60 feet or less.

If 40- and 80-meter DX capability is required and only space for one antenna is available, the 80-meter ground plane/40-meter 5/8-wave vertical will deliver the DX even *if* there *is* no room for 80-meter radials. If no 80-meter radials are used, loading against earth ground may improve results. This can be accomplished by running a number 6 or 8 wire from the common junction of the 40-meter radial system and the shield of the transmission line to the nearest earth ground.

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4. "The Radio Amateur's Handbook," 45th edition (1968), page 340, American Radio Relay League, Newington, Connecticut.
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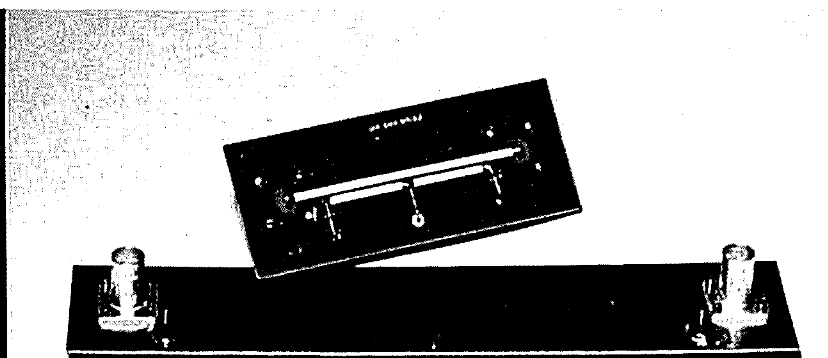
ham radio

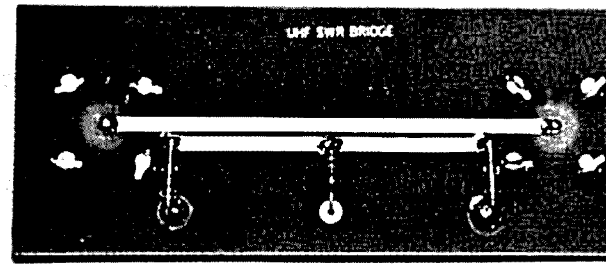


vhf and uhf micro-strip monimatch swr indicator

These two micro-strip monimatch swr indicators cover the complete frequency range from 40 to 600 MHz

Arthur R. Hall, W4CGC, Fairfax, Virginia 22030

This article describes the application of the micro-strip transmission line technique to an old favorite, in-line swr indicator, the *monimatch*[1]. Two models are described, a vhf model to cover the frequencies, 40 to 200 MHz, and a uhf model to cover the frequencies, 200 to 600 MHz. The original monimatch and



Construction details of the micro-strip uhf pickup line.

most of its decendents have been widely accepted down through the years as one of the easiest to build and most reliable of the in-line swr indicators.

As good as most of the monimatch designs are, they do have an upper frequency limit that prevents most of them from being reliable at vhf frequencies, and none seem to function reliably at uhf.

Actually, I tested the idea of adapting the micro-strip technique to the original monimatch design six years ago with a number of versions being built with good results. Since then, activity on the vhf and uhf bands has mushroomed tremendously, particularly on two-meter fm.

Most of the newer two-meter transceivers are all solid state and can be very unforgiving if they are used with an antenna that is not properly matched to the output circuit. The two micro-strip swr indicators described here will provide the means for reliably measuring swr and relative power in the region of 40 to 600 MHz.

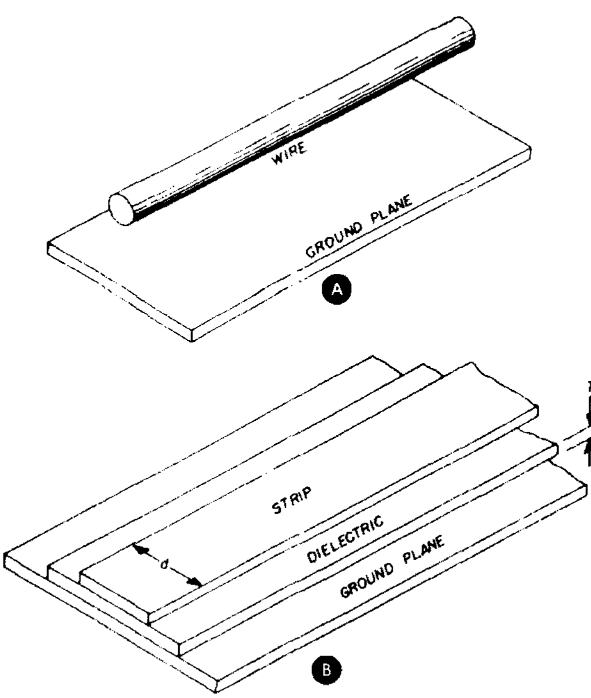


fig. 1. Cross section of the micro-strip transmission line system. Impedance is determined by the width of the strip, thickness of the dielectric and dielectric constant.

theory of operation

Basically, a micro-strip transmission line is an unbalanced, constant impedance conductor analogus to a wire above a ground plane with a dielectric in between as depicted in **fig. 1.** The characteristic impedance is dependent upon the line width, the distance of the line above the ground plane and the dielectric constant of the printed circuit material. (G-10 material was chosen because of its low loss factor at these frequencies, its wide

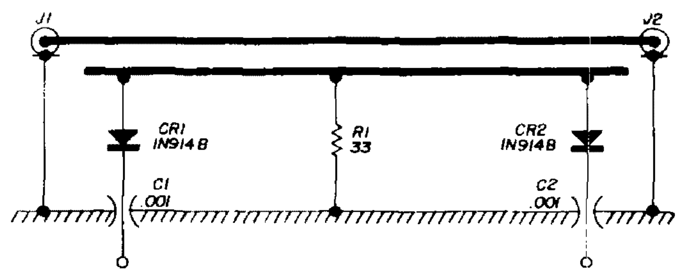


fig. 2. Circuit of the printed-circuit, micro-strip swr pickup line. C1 and C2 are Aerovox type 5601 discoidal feedthrough capacitors.

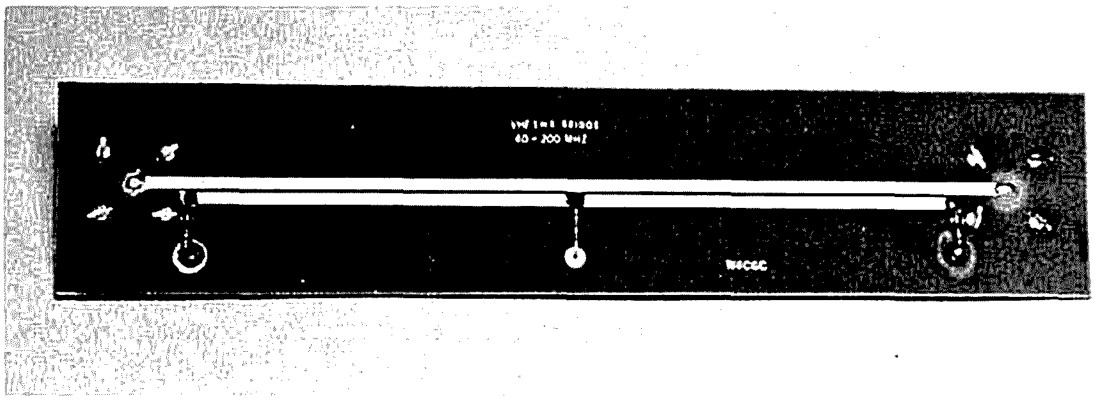
availability and its reasonable cost.)

A second line running parallel to the conductor line is center tapped to ground through a 33-ohm carbon resistor. High-frequency switching diodes rectify the induced rf voltage (on the pick-up line) which is then filtered by uhf feedthrough capacitors (see **fig. 2**). The pick-up line is coupled both inductively and capacitively to the main conductor line as shown in **fig. 3**. One half of the pick-up line samples the power flowing in the center line in one direction between one diode and the terminating resistance, the other half of the line senses the rf power flowing in the opposite direction between the other diode and the terminating resistance.

The upper frequency limit of these swr indicators is dependent not upon the microstrip center line, but upon the pick-up line. *It appears that the inductance* of the pick-up line from one end, to the terminating resistance and its distributed capacitance to the ground plane,

is the determining factor. One theory of limitation is that the pick-up line consists of multiple series inductances with multiple distributed capacitances to ground as illustrated in fig. 4A.

Another word of caution is in order at this point. Even though a micro-strip transmission line is designed very carefully to maintain a constant 50-ohm impedance, discontinuities may occur at each end of the micro-strip line where it connects to the coaxial connectors. It is for this reason that I have chosen to mount the two coaxial connectors directly to the micro-strip line. This method of attachment will maintain a near constant 50-ohm impedance between the coaxial connector and the line.



Closeup of the vhf circuit board shows the micro-strip pickup line, diodes and resistors.

Fig. 4A can be approximated as a simple pi network as in fig. 4B. Since a pi network is a low pass network, the cut-off frequency would then be a function of L2, C1 and C2. Laboratory measurements of these values correlate closely to the upper frequency limit of reliability of both the vhf and uhf swr indicators.

One important factor often overlooked in the design and fabrication of in-line swr indicators is that it *must* maintain a constant impedance, exactly the same as the characteristic impedance of the line it is measuring. If it does not, the in-line swr indicator can, itself, cause a mismatch even though it shows a perfect match between the generator and the load.

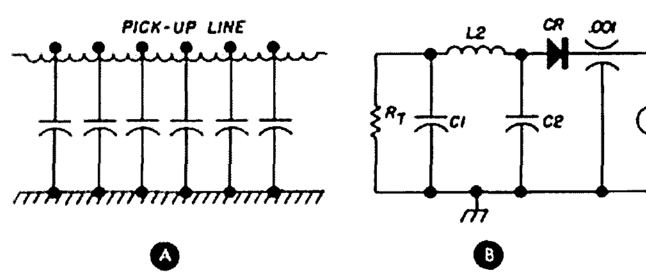


fig. 4. The distributed capacitance and inductance of the pickup line (A) limit the upper frequency limit of the swr indicator because they form the pi-network in (B).

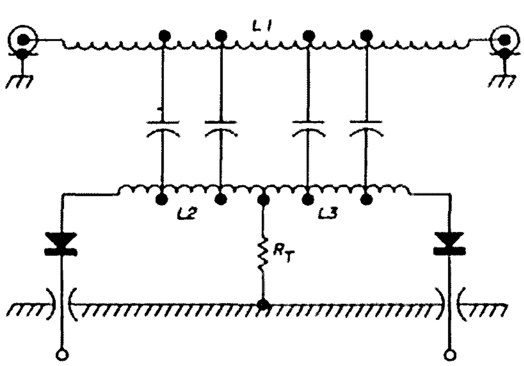


fig. 3. Upper operating frequency of swr pickup line is limited by the distributed capacitance and inductance of the line (see fig. 4).

Laboratory measurements of the inherent swr of the two micro-strips using a General Radio Model 900LB precision slotted line, is shown in fig. 5. The results are quite acceptable throughout the useful range of each indicator.

Another important factor in the design and fabrication of a reliable swr indicator is that it must be perfectly balanced between the center of both the conductor line and the pick-up line – one side must be a mirror image of the other. This is

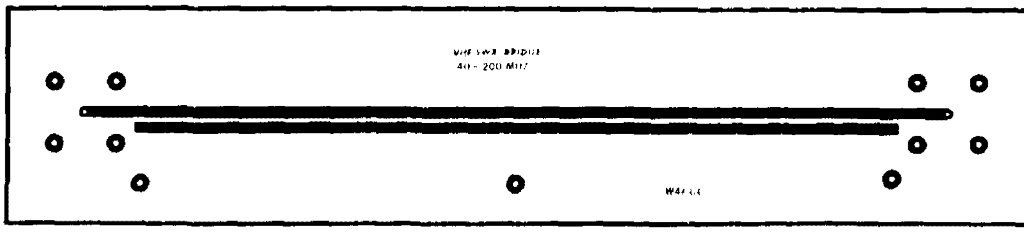
most easily achieved using the micro-strip technique.

construction details

As was pointed out earlier, accuracy of line width, thickness of dielectric and dielectric constant must be maintained to close tolerance for best results. Experiments show that when using G-10 epoxy-glass material of 1/16 inch thickness, a line width of .094" ± .002" will provide the necessary 50-ohm characteristic impedance. The copper-clad should be no thicker than 1-ounce material to minimize undercutting during the etching process. Both the vhf and uhf indicators follow the same construction technique; they are actually identical except for the line lengths.*

After the boards have been photo etched and trimmed to size they should be drilled using very sharp drills of proper size at high speed. The two holes that the center conductors of the connectors pass through should be beveled on the ground plane side so that a short does not occur. This beveling can be done with a counter-sink tool, or better still, a sheet metal 0.312-inch drill can be used with an appropriate pilot drill. The idea is to just go deep enough to remove the copper without cutting into the G-10 dielectric.

The two coaxial connectors are standard UG-290/U BNC units which have been modified for micro-strip use by being turned down on a lathe (see fig. 6). The face of the square flange should be faced off approximately .010" to remove any imperfections that might exist in the connector. When facing off the flange, continue all the way through to the center conductor, removing the shoulder boss. Then cut off the center conductor so that .080" remains. This will yield an rf connector that can be directly attached to the micro-strip printed-circuit board. Attach the modified UG-290/U connectors to the micro-strip board with ¼" long, 3-48 machine screws.



Half-size layout of the vhf printed-circuit board.

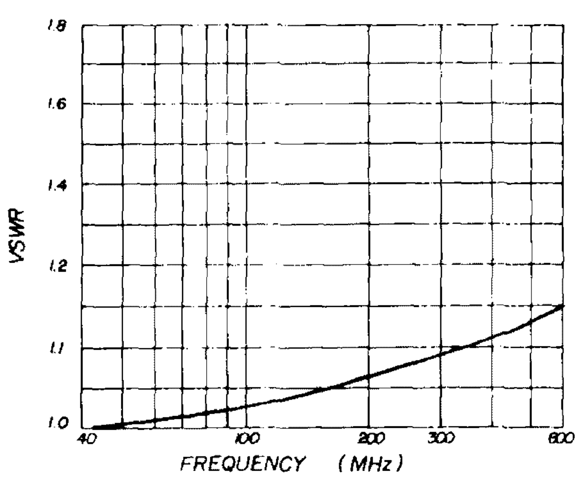


fig. 5. Insertion vswr of the two microstrip swr indicators.

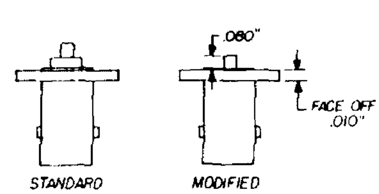


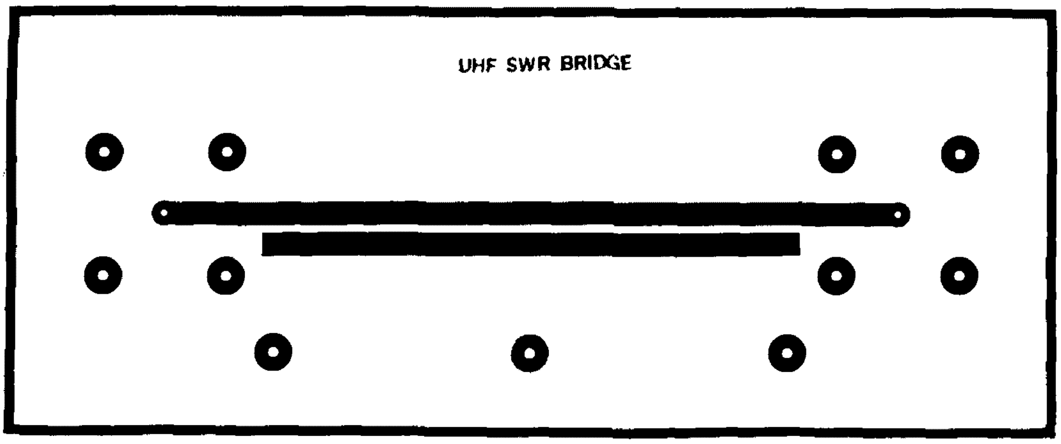
fig. 6. Modifying the UG-290/U BNC connector for use with the micro-strip line.

*Completely cut and drilled G-10 epoxy glass boards and complete kits are available from TRI-COM, Inc., 12216 Parklawn Drive, Rockville, Maryland 20852. Either the vhf or the uhf board is $7.00 and either complete kit is $13.50. Both prices include air-mail first-class shipping charges. Maryland residents, please include the 4% sales tax. Rush orders can be placed by telephone to 301-770-5585.

The next step is to sweat solder the two feedthrough capacitors in place. This operation should be done with a heavy duty soldering iron with a large tip. Make sure solder flows evenly all around the capacitor flange.

The soldering of the two center conductors of the modified UG290/U connectors should be done with the same heavy-duty iron. Use solder sparingly at these two points. This operation should be done quickly to reduce chances of peeling the foil from the board material. This is not a critical point but care should be exercised.

Lastly, install the two diodes and terminating resistor. No special care is needed here except to keep these components perpendicular to the pick-up line.



Full-size layout of the uhf micro-strip swr bridge.

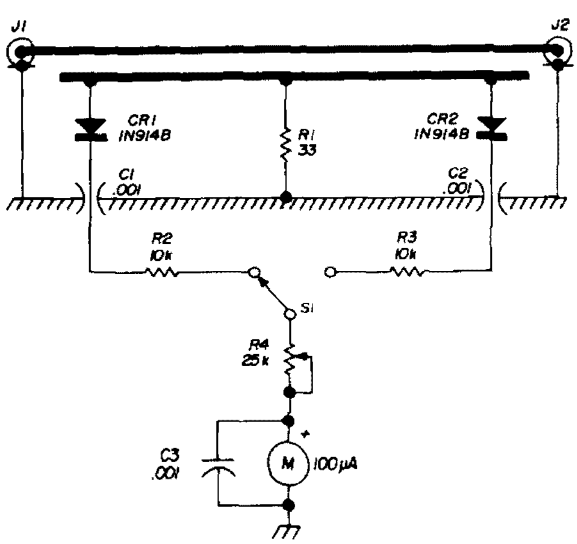


fig. 7. Complete circuit of the swr indicator, *including the micro-strip pickup* line and meter.

testing and calibration

The swr indicators can be tested for accuracy by inserting each indicator between an appropriate transmitter of 1-watt or more and a known 50-ohm *non-reactive* dummy load such as a Bird Termaline, etc. Most homebrew dummy loads, when used at vhf and uhf, will be reactive, thus producing a reflected reading on the indicators. In fact, these indicators should be accurate enough to indicate how good your dummy load is at these frequencies.

I have not made provision for enclosing the indicators in a mini-box, choosing to let the builder make his own decisions in this regard. The only word of caution here would be to keep at least a ½-inch clearance between the rear plane of the micro-strip and the enclosure. A suggested schematic including balancing resistors, forward-reflected switch, sensitivity potentiometer and meter is provided in **fig. 7.**

Acknowledgement goes to Mr. David W. Reynolds, W3QKR, for assisting with the circuit analysis, and to Mr. John Gregory, W3ATE, for his assistance in providing the photo-copy work.

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2. "Micro-Strip – A New Transmission Technique in the Kilomegacycle Range," *Proceedings of the IRE,* December, 1952.

ham radio

monitor receiver for RTTY autostart

A stable receiver for continuous monitoring of the RTTY autostart nets

Bert Kelley, K4EEU, Tampa, Florida 33609

This receiver was built to monitor the fixed-frequency autostart RTTY nets on the 80-, 40- and 20-meter amateur bands. If one RTTY station wants to leave a message for another, he sends it at a time when the band is expected to be open between the two stations. All of the stations in the net usually leave their equipment on 24 hours a day, and everybody copies all the messages transmitted. Of course, the addressee of a specific message doesn't have to be in attendance to receive a message. When 14.075 MHz is open, copy is excellent and results in a nationwide intercom system linking RTTY enthusiasts. Other nets on other bands are used for shorter distances.

overview

The equipment required for this type of operation is not elaborate. Many stations use ST-4 or ST-6 autostart demodu-

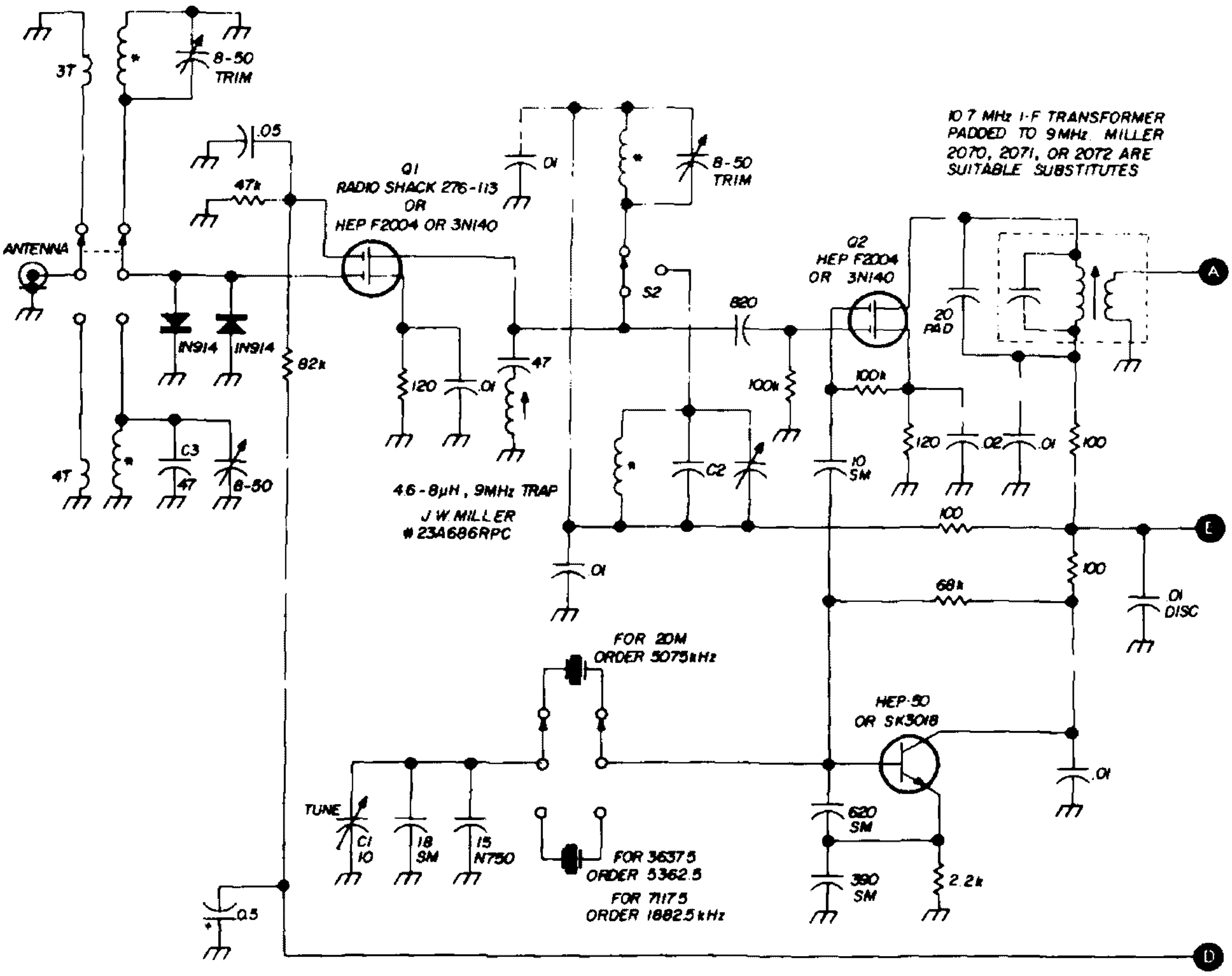


fig. 1. The fixed frequency monitoring receiver. C1 should be a panel-mounted variable. All the tuned circuits with the asterisks use the same inductance, all wound on Amidon T-50-2 forms. For 80 meters, the coils are 40 inches of no. 30 enameled wire; for 40 meters they are 29 inches of no. 26 enameled wire; and for 20 meters they are 13 inches of no. 24 enameled wire. Capacitors C2 and C3 are about 47 pF for 80 meters and are omitted on all other bands. Y1 is 5075 kHz for 20 meters, Y2 is 5362.5 kHz for 3637.5 kHz monitoring and 1882.5 kHz for monitoring 7117.5 kHz.

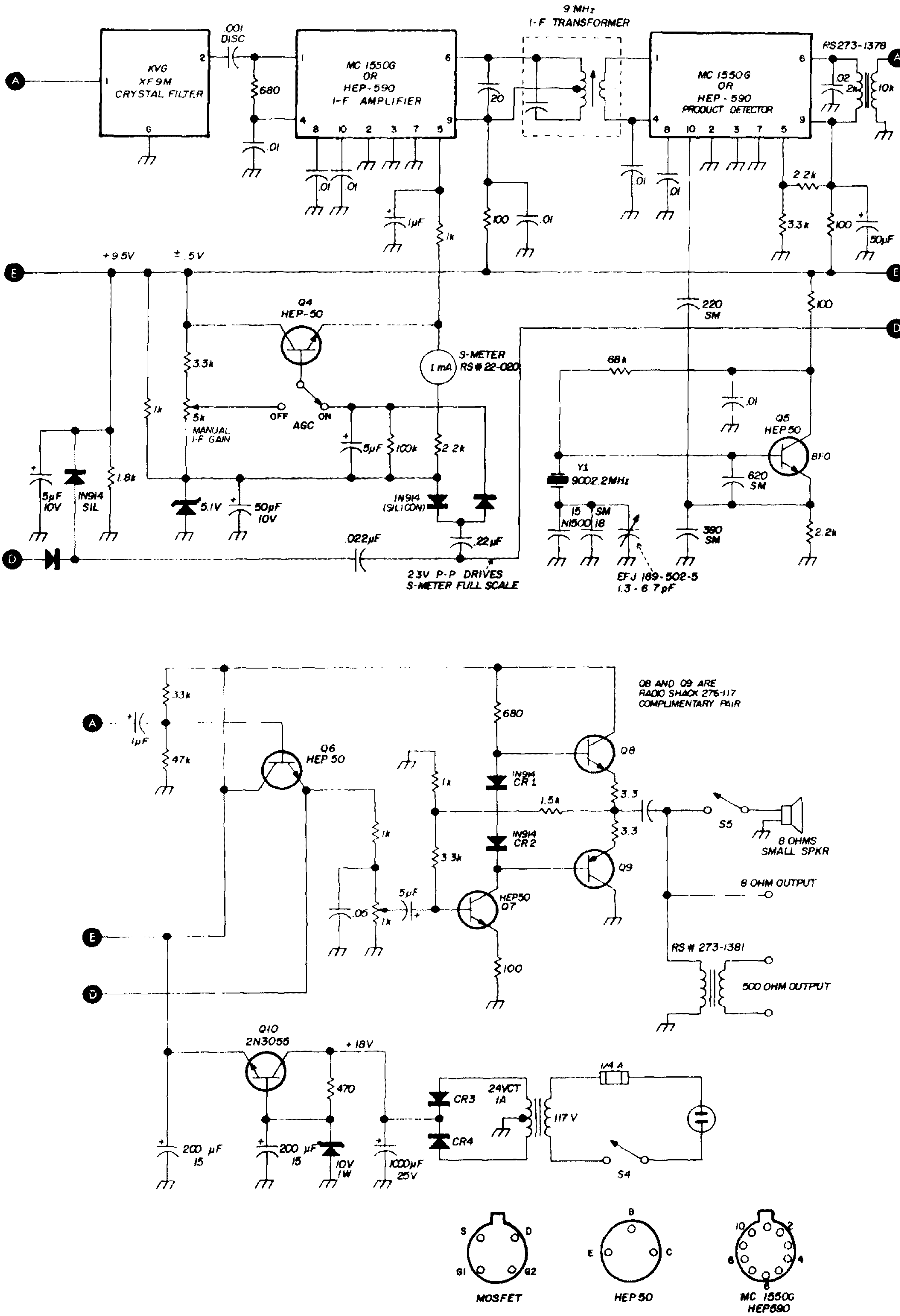
lators and Model 15 or Model 28 teleprinters.[1] Since only 170-Hz shift is used, it is necessary to operate close to the nominal frequency. Frequency precision and accuracy, therefore, are the most demanding requirements. Many stations have secondary-frequency standards and use quality amateur receivers with crystal-locked permeability-tuned *oscillators.*

A receiver that is to compete with such equipment must have good selectivity, sensitivity and *excellent* stability. The best way to get these features is to solid-state the old reliable superheterodyne circuit.

stability

The basic receiver design may be modified to fit other applications. The KVG filter is available in bandwidths that would be suitable for ssb, slow-scan television, 850-Hz shift RTTY or WWV reception. The front panel control provides a plus or minus 1-kHz tuning range. If a greater range is needed, a transistor vfo might be considered at a trade-off in stability.[2]

Crystal control is recommended because, properly compensated, the frequency drift of this receiver is less than 10 Hertz for a 20° ambient temperature change. To get this kind of stability, it is necessary to use a reasonably-priced high-accuracy crystal, specified for a 32-pF circuit capacitance and compensated for any remaining drift with properly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9 MHz I-F TRANSFORMER
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.02
2k
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1µF
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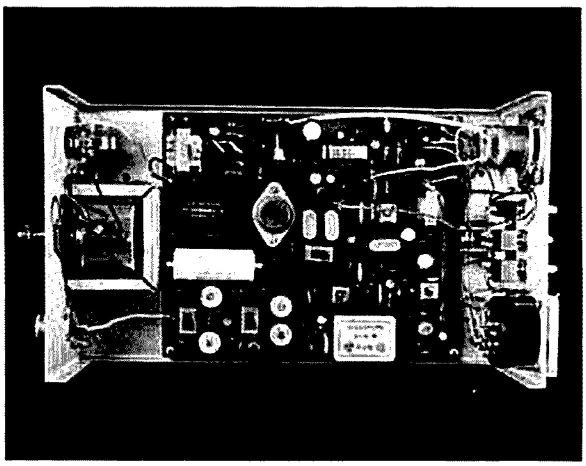
3.3k
50µF
+9.5V
+.5V
Q4 HEP-50
S-METER RS # 22-020
1 mA
220 SM
68k
3.3k
5k MANUAL I-F GAIN
OFF
ON
AGC
5µF
100k
2.2k
Q5 HEP50
BFO
Y1 9002.2MHz
620 SM
1.8k
5µF 10V
IN914 SIL
5.1V
50µF 10V
IN914 (SILICON)
22µF
.022µF
15 N1500
SM 18
390 SM
2.3V P-P DRIVES S-METER FULL SCALE
EFJ 189-502-5 1.3 - 6.7 pF
Q8 AND Q9 ARE RADIO SHACK 276-117 COMPLIMENTARY PAIR
33k
47k
Q6 HEP 50
680
IN914 CR1
IN914 CR2
1.5k
3.3
S5
8 OHMS SMALL SPKR
8 OHM OUTPUT
.05
HEP50 Q7
RS # 273-1381
500 OHM OUTPUT
Q10 2N3055
+18V
470
24VCT 1A
1/4 A
117 V
CR3
CR4
200 µF 15
10V 1W
1000µF 25V
S4
MOSFET
HEP 50
MC 1550G HEP590

selected N750 temperature-compensating capacitors. I used an International Crystal HA type crystal in my unit.*

The oscillator is a modified Colpitts circuit, and the large silver-mica capacitors effectively disconnect the crystal from the transistor junctions. The more capacitance that is used here, the better (up to the point where oscillation ceases).



Inside of the receiver. Antenna input is at lower left, audio output is at upper left. Bfo crystal is at right center. The two crystals mounted adjacent to one another are for 80 and 20 meters.

As might be expected, the 9002.2-kHz bfo crystal is the most temperature-sensitive component. While temperature compensating various crystals, I found that the ordinary garden-variety HC-6/U crystals drifted about five times more than the International Crystal HA type recommended here.

devices

After constructing a few receiver front ends using transistors and integrated circuits, it was a pleasant surprise to find how well the mosfet performed.[3] Forget the broadcast-band rejection filter and the attenuators in the antenna leads; mosfets provide plenty of gain without feedback. Toroid coils in the rf stage help, too.

The MC1550G IC is an inexpensive three-transistor array that works well as an agc controlled i-f amplifier and makes a terrific product detector.† The HEP590 is a similar device.[4]

agc

The audio-derived automatic gain control for the mosfet front end and IC i-f requires two different voltage levels and polarities. The mosfet requires a small negative voltage for agc, but the MC1550G agc becomes effective only after voltage rises to 5.1 volts.

After the voltage gain provided by the 2k:10k transformer, and impedance matching in transistor Q6, the peak-to-peak voltage at the emitter measures about 2.3 volts maximum, and averages about 1 volt. This ac voltage is rectified and charges a 5-μF capacitor, the ground end of which is connected to a 5.1-volt reference.

Another emitter follower, Q4, is used to give high-impedance input and low-impedance output. The emitter voltage follows the charge across the small capacitor and is used to drive the s-meter and provide agc to pin 5 of the i-f amplifier. Rf agc becomes effective after about two S-units.

audio

There is a variety of audio amplifier ICs available, but they are not recommended for this receiver because their high gain is not needed. Also, they usually will not stand sustained audio overloads. The transistor audio amplifier stages used in this receiver have the right amount of gain, are very rugged, and cost less than the IC. If 600-ohm output is not needed, no transformer is required.

construction

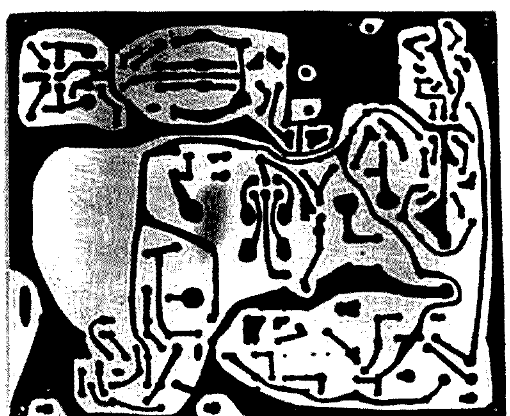
The 2N3055 used in the power supply is a rugged transistor often used in com-

*International Crystal Manufacturing Company, 10 North Lee, Oklahoma City, Oklahoma 73102. Write to them for their complete catalog with details on ordering high-accuracy crystals for your specific frequency and application.

†Both Hal Devices and Circuit Specialists stock them.

mercial equipment. You can find it surplus for less than $2. I recommend that you build the power supply and check it out first. When building the receiver, work backwards from the audio stages, checking stages as you go.

The photos show the receiver built on a 6-7/8 by 5½-inch circuit board mounted on spacers and inside a 10 by 6 by 3½-inch minibox. The board is laid out in a semicircle around the large electrolytic capacitors.



Printed-circuit board before etching. Antenna input is at top left, with mixer to right, and filter (large black area). Top to bottom on the right are the i-f, product detector and agc. Audio is at lower left.

In these days of kits and commercial equipment, it takes a special breed of amateur to build his own receiver. If you are a builder, and if you need a really good monitoring receiver, build this one. You'll like it.

references

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2. James R. Fisk, W1DTY, "Stable Transistor VFOs," *ham radio*, June, 1968, page 14.
3. John Knepler, "Cross-Modulation and Intermodulation in Receiver R. F. Amplifiers," *Electronics World*, March, 1970, page 55.
4. Motorola Application Note AN-247, Motorola Semiconductors, Phoenix, Arizona.

ham radio

building
Motran and Motrac
channel elements

Building simple plug-in crystal board replacements for the hard-to-find Motran and Motrac channel elements

Thomas McLaughlin, WB4NEX, St. Petersburg, Florida 33704

Finding channel elements for Motorola's popular Motrac and Motran fm transceivers can often be a problem. Getting crystals for these rigs is relatively easy, but channel elements are both scarce and expensive. You can get around the problem, however, by building your own receiver and transmit elements.

It is fairly easy to make an oscillator unit on a perf board and press it into service as a channel element by connecting it to the proper pins on the radio circuit board. The general circuit of a receiver element, fig. 2, consists of a crystal oscillator and resistor-diode com-

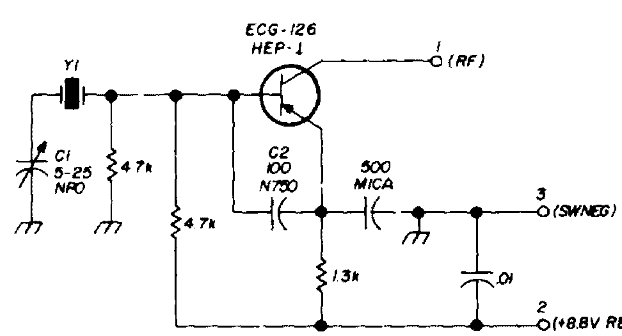


fig. 1. Simple receiver oscillator circuit. Precision grade crystals help to eliminate the need for the temperature-compensating components in fig. 1.

pensating network to correct for the effects of temperature change on the crystal. The circuitry for the transmitter element is the same except for the changes in the collector circuit as noted later. If you need a receiver element, build the circuit in **fig. 1** and order a *precision grade* crystal.

crystals

I have made several elements and ordered crystals from Sentry cut for circuit number 5 in their catalog. Each has tuned to frequency without problems. I found that a precision grade crystal will eliminate most drift problems even though the homebrew element does not have the compensating network. You will certainly stay within 0.001%, which is close enough for amateur work. This is exactly the same as buying a used element and installing the crystal in it yourself without adjusting the precision 1% resistors in the compensating network (fig. 1, labeled R) to eliminate oscillator drift with temperature change. These resistors may be anywhere from 9k to 15k depending on the particular characteristics of the crystal involved, and it really isn't worth the time it would take to find the correct values unless you have access to some expensive test equipment. If you need a transmitter element, build the circuit in **fig. 3.** This is exactly the same as **fig.** 2 except the collector of Q1 is grounded to pin 3 and the rf is taken directly from the emitter. **Fig. 4** is a guide to finding the proper pins.

If you have a used element or home-brew oscillator which won't tune on frequency, pad C1 with a 10-pF 10% NPO ceramic disc. This will work if the frequency is too high. If the crystal is too low in frequency or if no amount of padding helps, change C2 from 100-pF N750 to 75-pF N750 disc ceramic. This decreases the circuit capacitance enough to raise the frequency a little. Note that the new style single oscillator elements have holes in the circuit board so you can pad C1 if necessary. The older dual oscillator elements don't have this provision, but C1 is a 5 to 25 pF NPO unit so it shouldn't need padding. The new elements have a 3.5 to 14 pF N300 unit for C1 which accounts for the padding provision. Other than this slight difference, the circuitry for both the old and new types is the same.

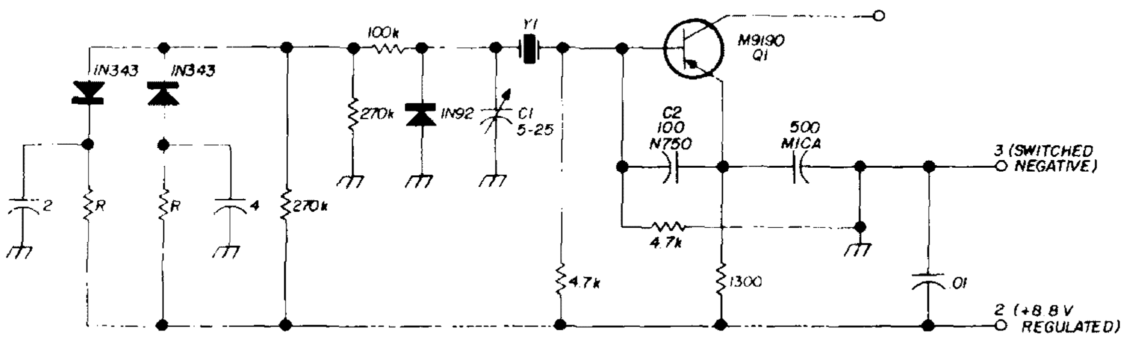


fig. 2. Basic receiver channel element circuit. The text explains about the resistors marked R.

If you have regular elements and are ordering crystals, it is best to get precision grade crystals cut for the particular model of element you are using. **Table 1** gives the proper designation for many elements. If you are ordering crystals for home-built elements, I would advise specifying them to be cut for element model

table 1. Motorola channel element designation.

New style single-frequency element

	0.0005%	0.0002%
transmitter	TLN 1083	TLN 1087
receiver	TLN 1081	TLN 1086

Old style two-frequency element (all are 0.0005%)

	single freq	dual freq
transmitter	TLN 1024	TLN 1025
receiver	TLN 1020	TLN 1021

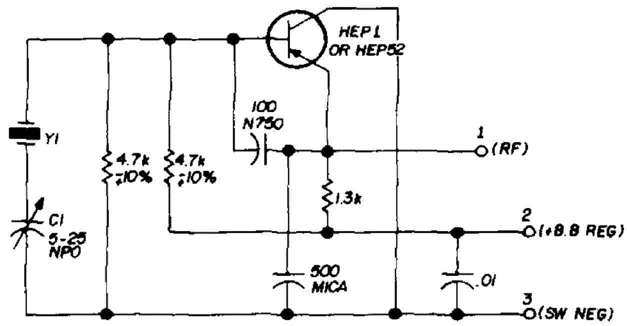


fig. 3. Transmitter oscillator circuit.

TLN1081, or if you buy from Sentry, specify circuit 5 in their catalog. The crystals should tune without any problems in the homebrew circuits.

The ersatz elements I made are all on perf boards, and I mounted them in the rigs using the pins from an old octal socket — they just fit on the male pins on the radio circuit board. If you're desperate or just don't care, you can even solder the element right into the radio.

transistors

The transistors in the standard elements are Motorola type M9190, a house number only. Since this is a PNP rf type, HEP-1 works well, but it is no longer being produced and is not available at all distributors. HEP-52 works well in transmitter elements, but the output is a little too low for use in a receiver element whose output is run right into a tripler stage, unlike the transmitter exciter. Sylvania ECG-126 is also a good bet for either receive or transmit elements because these are fairly high-output devices.

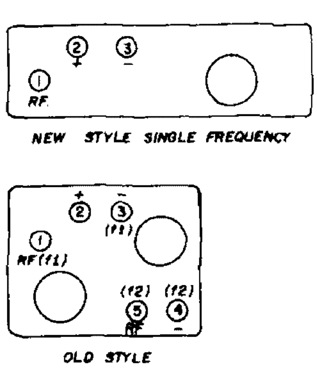


fig. 4. Bottom view of old and new Motorola channel elements shows pin designations.

There are countless other types which will work; even some audio types will have high enough output to be used.

If you have one of the older two-frequency elements which has only one oscillator board in the can, it can easily be converted into a functioning two-frequency element if you have a 5 to 25 pF NPO trimmer and oscillator board out of an unused element. Likewise, any receiver element can be changed to a transmitter element and vice versa. This only requires connecting the wire from the rf terminal to the proper transistor lead on the board and either grounding or ungrounding the collector as the situation dictates (see fig.

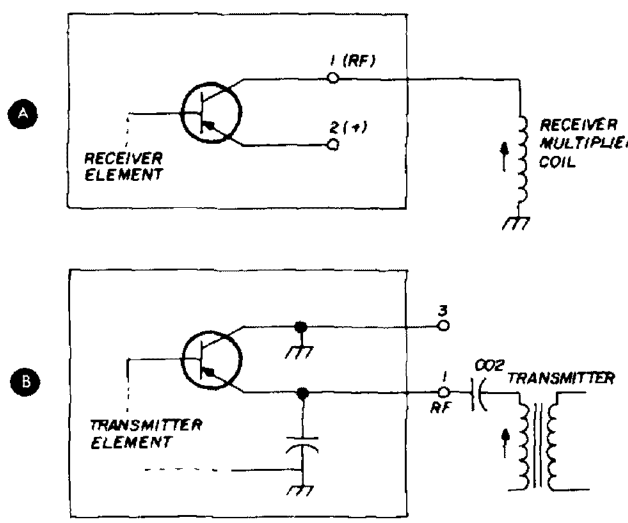


fig. 5. Difference in the hookup arrangements for a receiver element (A) and a transmitter element (B).

5.). If you haven't noticed, receiver and transmitter elements are not interchangeable.

birdies

Another characteristic of Motracs and Motrans is that the i-f crystal is usually 8.455 MHz, Motorola part G09. If the receiver is set up to receive on 146.94 MHz an intermod type birdie will appear in the receiver. This condition can easily be eliminated by changing the i-f crystal to 7.545 MHz, Motorola part G11, available either from Motorola, Sentry or International.

ham radio

monitoring oscillator

The step-by-step evolution and modernization of a very handy station accessory

R. W. Gunderson, W2JIO, 980 Waring Avenue, Bronx, New York 10469

However sophisticated some of our modern transceivers may be, sometimes they can benefit from some old tricks. The monitoring oscillator, for instance, has been around for decades but it still can serve very useful functions. I feel the modernization of this handy gadget might be of interest to old-timers who used one in the 1930's, to newcomers who probably never heard of the critter and to experimenters who might enjoy sharing my adventures in adapting modern components to an old circuit.

When you operate your transceiver on CW, you have a so-called side-tone oscillator to monitor your sending. However, you can't monitor your signal as it sounds to the distant station unless you have an additional receiver. If something goes awry while you're on the air – your final takes off, your frequency shifts suddenly or the note goes sour – you will never know about the trouble until the operator at the other end of the circuit tells you that something has happened. Further, the regulations specify that the frequency of the station must be measured by external means other than the station receiver. When you speak of frequency measurements, you mean the operating frequency, of course, and the *quality* of the emitted signal – since its frequency will be shifted if ripple appears in the output or if the stage takes off due to amplifier instability or loss of excitation.

The side-tone oscillator operates at just one frequency (usually about 1 kHz). I like to shift the frequency of the monitoring oscillator as I operate to break the monotony. Of course, this is impossible with the built-in side-tone generator. Rf-actuated monitors which derive their dc power from the radiated energy from the transmitting antenna are satisfactory. However, they do not check the quality of the emitted signal as it sounds to the distant operator.

the monitoring oscillator

In the days when amateur transmitters employed self-excited oscillators to generate the carrier frequency, a monitoring oscillator was an invaluable tuning and operating aid. The monitor is a simple oscillating detector enclosed in a shielded container. The shielding guarantees that the signals picked up on this receiver from the output of your transmitter will be weak, and will not overload the monitor. The typical instrument is made with a minimum of components and is fitted with plug-in coils to make band changing easy. My original pre-war monitor was made in a metal lunch box and it operated from 160 through 20 meters by means of four properly wound plug-in coils. The other day I dug it out of the junk box and modified it by replacing the 1G4 triode oscillating detector tube with a 3N139 field effect transistor. I've been using it ever since, and it does such a good job that I thought the monitor and its applications might make interesting reading for other amateurs, newcomer and old-timer alike.

The circuit for the original monitor as it was constructed here at W2JIO is shown in fig. 1. S1 was part of the "filament lighting" type phone jack. You will note that the earphones are in the negative (ground) lead of the detector. This is done to keep hand capacity effects to a minimum, improving the frequency stability. Pickup from the phone cord is also minimized, and the unit really performs as a weak-signal receiver.

The unit is used to set the amateur transmitter within an amateur band in the following manner: Set the station receiver inside the amateur band. This may be checked by means of the receiver's dial calibration and by listening to other stations in the particular band of frequencies. Tune the monitor to zero beat with the receiver frequency by turning on the receiver bfo and then setting the monitor to zero beat with the receiver. Disable the receiver by placing it in the stand-by position and listen to the monitoring oscillator. Then tune your transmitter to give zero beat with the signal in the oscillating monitor. The transmitter frequency is now set close to the original frequency tuned in on your station receiver.

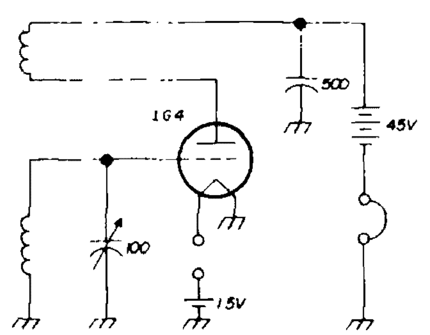


fig. 1. The original monitoring oscillator used a 1G4 triode and plug-in coils to cover 160 through 20 meters. The unit was turned on by a filament lighting type headphone jack (not shown here for clarity).

In the case of a radio-telephone transmitter operating on a-m, the quality of the carrier may be checked by listening to a beat note produced between the monitor and the carrier frequency, and then modulating the transmitter. If the pitch of the signal varies during modulation, there is frequency modulation of the a-m carrier rather than just amplitude modulation. In the case of your CW transmitter, tune the monitor to give a beat note with the transmitted signal and monitor the keying characteristics just as you might monitor them in a modern receiver. The overload characteristics of the monitor

are relatively good because the entire monitor circuitry is shielded, and the character of the signal may be judged just as it appears to the distant receiving station.

When most amateurs used regenerative receivers, the monitor was a must in the operation of a station. Today, however, the stability of the average communications receiver is good enough so that the transmitter may be set directly to the operating frequency. In the case of transceivers, there is no problem because the transmitter is automatically set to the receiver's frequency.

board to accomodate the leads for the fet, and wired it in place of the tube, just as in the original circuit. The transistor oscillated right off the bat, and I could hear the beat note from my transmitter – the job was finished – or was it?

When I pulled the coil out of the socket to try the monitor on 20 meters, I found that the transistor had gone bad. I had read that you have to be careful with these insulated gate fets, and I hadn't been careful. That gate must remain at ground potential or random charges will simply burn out the microscopically thin insulation. I simply connected a one-meg-

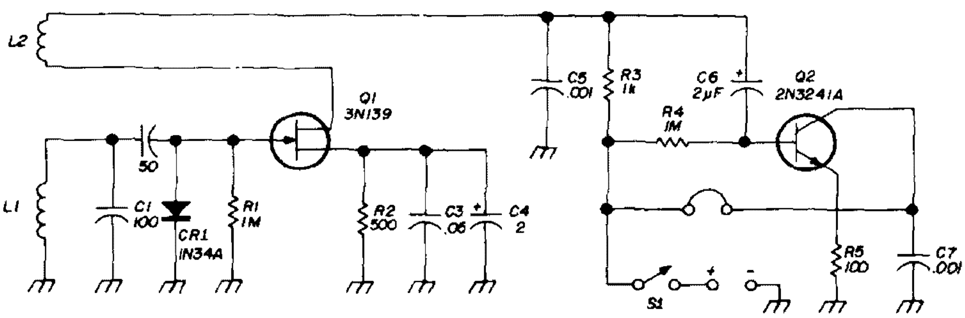


fig. 2. The modernized monitoring oscillator. Coil-winding data on the L1-L2 plug-in coils is given in table 1.

modernization

The final, updated schematic of the monitoring oscillator is shown in fig. 2. The steps to get to this point, however, were many. When I dug the old monitor out of the junk pile, I decided that the gadget would have to be transistorized. The tube filament was 1.4 volts at 50 mA, and the B supply was a 45 volt battery. Both batteries were expensive, and if I was to use the monitor as much as I expected, batteries would become costly. My first thought was simply to replace the 1G4 triode with a field effect transistor, a type 3N139 which I happened to have on hand.

I removed the tube socket and set a small piece of perforated board in its place – held with the same machine screws and nuts which held the original socket. I set four eyelet type terminals in a square through the perforations in the

ohm resistor from gate to ground, and found that I could change the coil without fear of destroying the fet.

I tried the circuit with my surplus HS33 headphones, which have an impedance of 500 ohms, and decided that perhaps high-Z phones would give me slightly more audio to make monitoring easier. However, upon plugging the 2,000-ohm phones into the unit, I found that the transistor didn't oscillate. In fact, the total current from the battery had fallen to practically zero. Somewhere back in the distant past, I had read that a source resistance would tend to make the transistors less critical as to the load, and that transistors of the same type could be plugged into the socket and the circuit performance would be more uniform. A 500-ohm resistance shunted by the usual 0.05-mF disc ceramic capacitor from source to ground did the trick. I could now use the circuit with both headsets,

and the output was even higher with the source resistor in the circuit.

audio stage

With a supply of 9 volts, I found that the output in the headphones wasn't quite what it was with the tube operating from the 45-volt B battery. Therefore, I decided that a single grounded-emitter stage as an audio amplifier following the oscillating monitor would be a worthwhile addition. This makes a very handy monitor — one which can now be used to check for harmonics as was the older circuit back in the 30's. The original

table 1. Coil-winding data for the updated monitoring oscillator. All coils are wound on standard 1¼-inch, four prong forms. Each coil has an approximate frequency ratio of three to one. Approximate frequency coverage is then: 160 meters from 1.75 to 5.10 MHz, 80 meters from 3.4 to 10.5 MHz, 40 meters from 7.0 to 21.5 MHz and 20 meters from 12 to 30 MHz.

160 meters	L1, 50 t. no. 28 enameled wire closewound; L2, 6 t. no. 28 enameled wire.
80 meters	L1, 34 t. no. 24 enameled wire, closewound; L2, 4 t. no. 24 enameled wire.
40 meters	L1, 16 t. no. 24 enameled, space wound to diameter of wire. L2, 3 t. no. 24 enameled wire.
20 meters	L1, 6 t. no. 24 enameled space wound at about twice the wire diameter. L2, 2 t. no. 24 enameled wire.

In all cases, L2 is closewound at the cold end of L1, in the same direction as L1.

phone jack with the additional filament switching leaves is still used, although it was necessary to insulate it from the chassis because the jack is now in the hot side of the circuit. As these jacks are scarce, you might prefer using a plain jack and simple SPST switch. This switch and the old bakelite vernier tuning dial are the only panel controls.

I found that the circuit starts more readily at the higher frequencies if the tank circuit is isolated from the gate of the fet by means of a small coupling capacitor of about 50 to 100 pF. The current drain of the oscillator is about 5 mA with the circuit oscillating, and this current falls to about 2.5 mA when the circuit falls out of oscillation (as produced by placing your hand on the tuning capacitor stator). This effect is just the opposite of that occurring in a vacuum tube oscillator and I decided to look into it before completing the design of the instrument.

As the gate is completely insulated from the rest of the field effect transistor, there can be absolutely no rectified dc flowing from the gate to the source through the gate leak resistor, R1. Therefore, there will be no dc voltage developed across this resistor to provide additional operating bias during oscillation of the circuitry. I decided to furnish the necessary bias by connecting a 1N34A small-signal germanium diode with its anode to the gate and its cathode grounded.

With a supply of 9 volts applied to the oscillator drain, the negative voltage is now about minus 3 volts from gate to ground. The drain current now decreases when the circuit oscillates, just as it does with the vacuum tube oscillator, and the audio output from the monitor has increased markedly with the introduction of the diode. It is quite possible that a higher value of source resistance from the fet source to ground would enhance the weak-signal performance of the instrument. However, it is quite satisfactory with the values shown in the circuit.

The monitor performs very well all the way from 160 to 10 meters, by inserting the proper plug-in-coil. It will oscillate at frequencies higher than 30 megacycles, although the oscillator stability is rather poor at these higher frequencies. This instability is produced by the poor mechanical arrangement of components and the relatively large amount of slip in the old bakelite vernier dial. However, it is reasonably satisfactory — good enough for monitoring my CW.

ham radio

helically wound mobile antenna

Improved performance over a manufactured whip is claimed for this antenna design

H. L. Booth, ZE6JP, Salisbury, Rhodesia

This article describes a helically wound whip antenna for mobile operation. The final design evolved over a period of about five years. An antenna was desired that performed better than those available on the market; tests have indicated that this objective has been achieved.*

The antenna has flat response at resonance and frequencies above resonance, with pronounced fall-off at frequencies below the design frequency.

Design data and construction details are given to enable you to duplicate the antenna, either as a single-band design (1-150 MHz) or as a 4-band amateur antenna covering 10-80 meters.

Construction procedures, dimensions, and winding instructions must be followed explicitly, otherwise the antenna may not perform as claimed. After you've built the antenna from the instructions provided here, *then* try your own variations. But it's important to "stick to the script" to start with.

*A copy of the test report is available from *ham radio* for *$1.00* and a self-addressed stamped envelope. editor.

During the 1964/65 period, I conceived an idea to build a single-band helically wound whip into a two-band antenna while also trying to improve the coupling to space by the production of a near-sinusoidal current and voltage distribution over a short antenna. The results have been very satisfactory.

Having made many single-band helically wound whips, I noticed that a second resonance was apparent around 18-19 MHz on most antennas, using rod about 3/8 inch in diameter for the dielectric. While developing the technique to wind single-band whips for frequencies from 3.5 to over 100 MHz, trends were noted, and an antenna for the 40- and 20-meter bands was attempted. My first attempt, which was pure luck, was a helically wound antenna similar to the present multiband antenna. After rewinding and making adjustments, the first design was born. It worked on 40 and 20 meters, so I tried it on 10. It loaded and worked, but as this was done during the quiet period of 10 meters, only local results were obtained. Finally the antenna was tried on 15. It worked on that band also. Subsequent results have been most satisfactory. Tests on 20 showed 3 dB gain over a Hustler at a distance of 14,500 miles.

The form factor was a compromise, producing a near-sinusoidal distribution of voltage and current similar to that of a ¼-wave atnenna.

single-band design

Experimentation has resulted in a formula for determining the approximate

length of wire for a helically wound antenna for one frequency. The formula is at best an approximation as shape, dielectric rod length, and wire gage affect the formula. To find the approximate length of wire for a helically wound whip for one frequency, use

$$L = \frac{840}{F}$$

where

L = wire length (ft)

F = frequency (MHz)

This formula will result in a little more wire than required providing the top third of the antenna length is close-wound. If less than one-third is close-wound, more wire will be required; conversely, if more than a third is close-wound less wire will be required.

The dielectric rod must be of constant diameter. Tapered rods will result in a different configuration than that specified, which may affect performance.

The rod length represents a quarter wavelength or 90 electrical degrees. Divide the rod length into nine sections, each of which represents 10 electrical degrees. To find the percentage of turns required at each 10-degree segment, use the data in fig. 1.

wire gage and length

Consider now one-third of the rod length. From fig. 1 note that 71 percent of the total turns, or wire length, (using a constant-diameter rod) must occupy that space. From geometry the rod circumference is πD. Therefore, dividing 71 percent of the wire length by the rod circumference will give the approximate number of turns to be closewound. From the wire table in the Handbook, find a suitable gage of enamelled wire. The wire diameter should not be less than 0.028 inch.* If the wire table gives a size smaller than 0.028 inch, use a larger-diameter rod and recalculate.

Helically wound mobile whip for 10-80 meters used by ZE6JP.

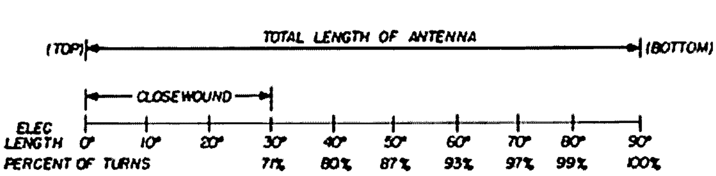


fig. 1. Diagram showing percentage of wire turns as a function of antenna length in electrical degrees. Approximately 70 percent of the total number of turns must be close-wound over the top one-third of the antenna.

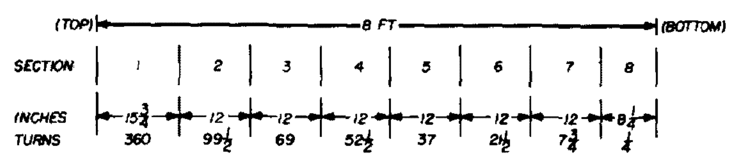
Using the formula above, a 3/16-inch-diameter rod, 18 inches long, was used to build an antenna for 10 meters. The antenna was mounted on the car and tuned. An input of 22 watts was used. Good reports were received across town, but after one minute of operation the antenna was too hot to touch because the wire gage was too small.

winding procedure

Mark off the rod into 9 sections. It will be easy to determine the number of turns in each section as the rod circumference is known; also the total length of wire. Divide the circumference into the length to obtain the total number of turns. Divide each section into inches. Note that a change of turns per inch

*The ARRL Handbook shows this wire diameter as No. 21 B&S gage. The current-carrying capacity of No. 21 B&S gage, at 1500 circular mils/ampere, is 0.54 amp. editor.

exists, section-to-section. Mark each number of turns in each progressive inch to accommodate the change of turns per inch. The winding then will have a constant change in pitch, and no sudden change of pitch will be obvious. Indelibly impedance feed conditions at resonance are not normally harmonically related; however, this antenna does have this property. The resonances occur in the ham bands, and the feed impedance allows the antenna to be loaded by the



SECTION	WINDING DATA	
1	360 TURNS CLOSE WOUND	
2	[illegible]	(TURNS/IN.)
3	[illegible]	(TURNS/IN.)
4	[illegible]	(TURNS/IN.)
5	[illegible]	(TURNS/2 IN.)
6	[illegible]	(TURNS/2 IN.)
7	[illegible]	(TURNS/2 IN.)
8	¼ TURN OVER LAST 8¼ IN.	

fig. 2. Winding details for a 4-band amateur antenna. No. 18 AWG enamelled wire is recommended.

mark the position of each turn. Anchor one end of the wire, then secure the other end to the "loose-wound" end of the fiberglass rod and wind on. After completing the winding lock the top end with tape, then adjust the turns to smooth out any uneveness in the winding. Secure the entire winding with epoxy.

Mount the antenna in its operational position. Make certain the car has an open space of at least 20 feet around it. Use a two-turn loop to ground the bottom end of the antenna. Couple a gdo to the two-turn loop. Check the gdo frequency with an accurately calibrated receiver. The frequency should initially be lower than that required. Remove turns from the close-wound (top) end, turn-by-turn, until the gdo dips at the low end of the band. The antenna will load over the band by adjusting the transmitter tank circuit.

multiband design

This is an extension of the single-band design, but by its size and shape it will operate satisfactorily on the 40-10 meter bands. The multiband version behaves like an hf choke. As frequency is increased, resonances occur at different frequencies. These resonances are governed by the antenna shape, wire inductance, and distributed capacity. The low-conventional mobile pi-section tank circuit. By adding approximately 60 μH in series with the antenna base most of the 80-meter band may be covered. Similarly, the antenna will tune 160 meters with suitable inductance added at the base.

From band-to-band, the feed-point impedance at resonance varies but is generally between 15-50 ohms. No difficulty has been experienced when feeding with RG8/U about 15 feet long. Forget swr so

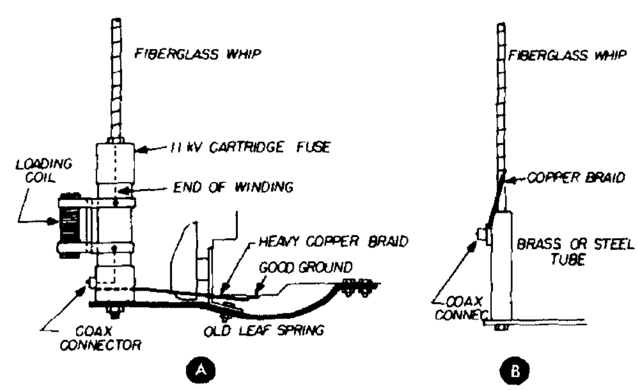


fig. 3. Suggested mounting arrangements. A shows author's mount, which includes loading inductance. A simplified version is shown in B.

long as the antenna can be loaded – improving swr adds very little to the radiation. In general, the usual pi section is adequate, unless in certain manufactured transceivers the 50-ohm termination is restricted.

The rod is of fiberglass with a diameter of ¾ inch, from 8 to 8 ft. 3 in. long. Lay the rod on a bench or table. Mark off the turn positions along the rod (fig. 2). Scratch the marks so they won't rub off when winding. Mark off from the top end as in fig. 2. Wind as previously instructed and terminate in the same way. Use only 0.040-inch enamelled wire (18 AWG). A suggested antenna mounting is given in fig. 3.

tuning

All previous instructions apply except as follows. Remove turns from the top of the antenna until it resonates in the low end of the 40-meter band. Check resonances on the other bands with the gdo. An increase or decrease in rod diameter will change the resonant frequency.

Note that after each adjustment of the antenna a check over the band, on each band, should be made. A compromise may be necessary in some cases, but this was not found to be so in my experiments. While testing, the antenna must be in its normal operating position. Changing from mobile to mobile may require some readjustments. Removing turns from the top has a profound effect on 40 and a lesser effect on 15; less still on 20 and 10.

80- and 160-meter operation

The antenna may be used for the two lower amateur bands by adding a suitable loading coil. The antenna should be resonated, as before, at the lowest frequency of the band you intend to use. Again, check loading across the entire band. The antenna should take power if your transmitter output circuit is not too restricted.

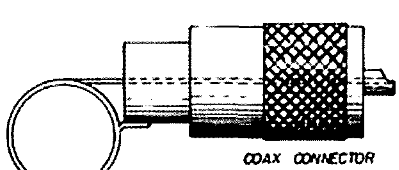


fig. 4. Loop with coax plug can be coupled to grid-dip oscillator to check resonant frequency of the antenna.

conclusion

I am informed by ZS6U that he has designed a 40, 20 and 15-meter single section, which screws onto a Hustler in place of the loading coil. Changes of wire gage are used for this antenna, but details are not available. Performance is at least equal to the single-band arrangements.

Close-up of ZE6JP's mount. Details are shown in fig. 3.

Additional resonances have been noted but no attempts have been made to use them. Typical resonances are (in MHz): 3.62, 7.05, 14.2, 21.1, 28.28, 31.8, 37.42, 44.5, 56, 67, etc. I don't know what the polarization really is, except it is mainly vertical by response on vhf. There is less decrease in signal strength when the antenna is moved from vertical than that measured from a base-loaded vertical antenna under the same conditions.

It's nice to change bands inside the mobile simply by reloading or by switching a relay to remove the short across the 80-meter coil.

ham radio

simplified autostart and antispace for your ST-5

This simple circuit can be easily added to your ST-5 RTTY demodulator to provide both autostart and antispace operation

Joseph M. Hood, K2YAH, 67 Mountain Ash Drive, Rochester, New York

It wasn't long after getting my teletype station operating that I became annoyed at having to constantly switch my teleprinter off, on and into *mark* when tuning or following a transmission. Before getting the teletype on the air the concept of autostart and antispace circuitry seemed to be a luxury that I could do without. However, after a few days of operating, it fast became a necessity. So began the search for suitable circuitry that would put an end to my switch-throwing frenzy.

I use an ST-5 terminal unit, fathered by W6FFC, which has been described sufficiently in previous articles. Autostart and antispace circuitry for the ST-5 has been also described previously, but every circuit which I happened upon seemed much too complex for the task at hand. So I decided to come up with something on my own – a simplified autostart, antispace (SAA) circuit.

Both autostart and antispace functions require a level-sensing device with a fairly high input impedance and good current handling capability in its output stage. After a short perusal through the integrated circuits catalogs I came across the Amperex TAA560 level detector Schmidt trigger. Its characteristics include

high input impedance, low power requirements and high output current control. It looked like just the thing for a start on a simple autostart and antispace circuit.

autostart operation

The TAA560 is a four-terminal device *and its operation is straightforward.* A TAA560 is used to detect the presence of a signal for the front-end portion of the autostart circuitry. However, to complete the circuit it is necessary to build a time-delay network for use in the teleprinter on-off control portion of the autostart circuit and in the antispace circuitry. Again, the TAA560 is simple to

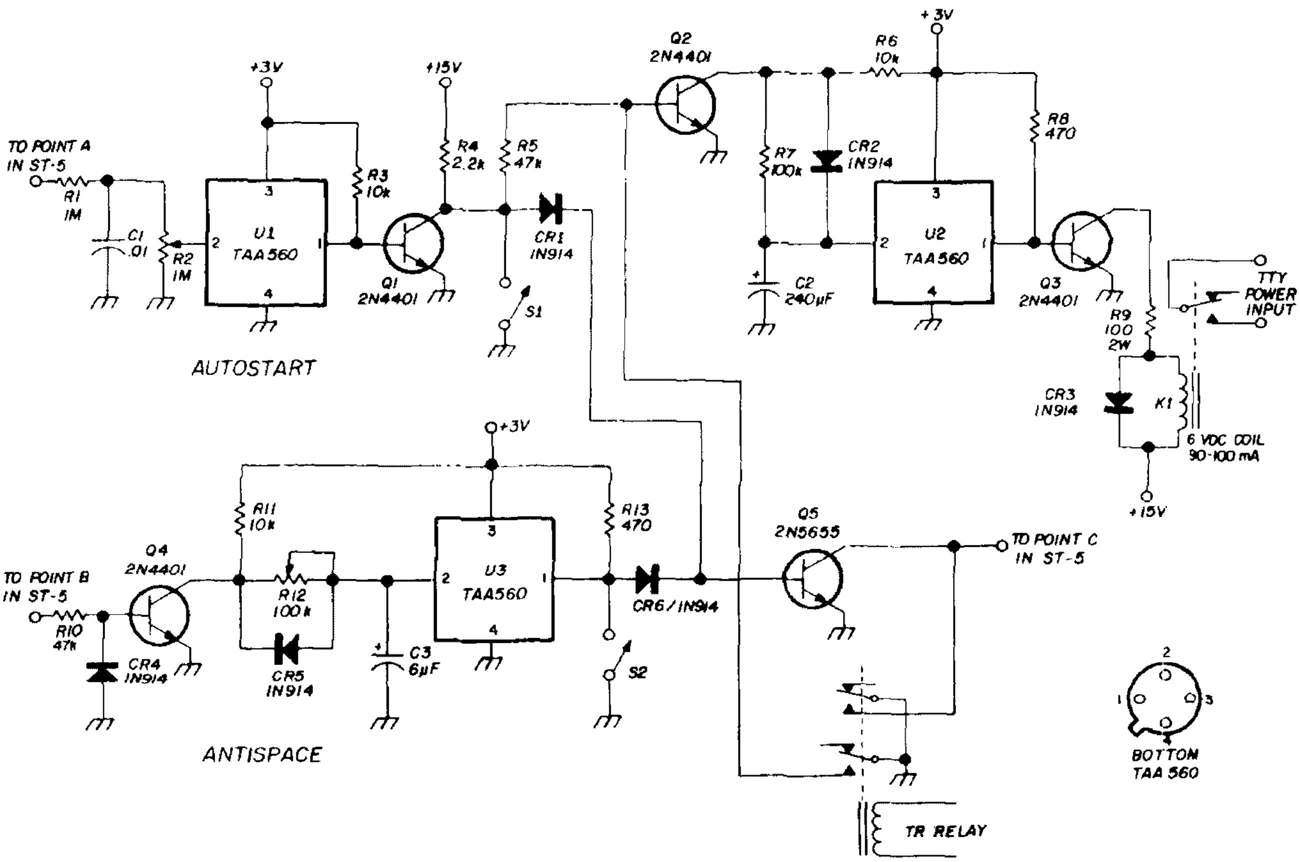


fig. 1. Complete diagram of the simple autostart, antispace circuit for the ST-5 RTTY demodulator. K1 is a 6-Vdc relay, spst.

supply voltage of 2.5 to 4.5 volts is connected to pin 3, and pin 4 is connected to return. If the input signal applied between pins 2 and 4 is above the circuit trip point (between 1 and 1.5 volts) the output transistor will be off, and the pin 1 to pin 4 output path is open circuited. However, once the input level drops below the circuit trip point, the output transistor in the TAA560 is biased on, and the pin 1 to pin 4 path will pass up to 50 mA of current.

This level-detecting feature of the use. A time delay may be had by using an RC network in association with the TAA560 input circuit. The capacitor is tied directly across the input of the TAA560 and is charged through a series resistor.

When a dc level is applied to the free end of the charging resistor the circuit waits until the capacitor is charged to the trip point of the TAA560 before changing state. The resistor-capacitor time constant determines the resultant delay. If a different turn-on versus turn-off delay is

required, a diode can be used to switch a resistor into or out of the charge or discharge path, as required.

With the level detector and time delay *functions in mind, let's look* at the total circuit operation by referring to **fig. 1**. Start with the autostart portion of the circuit.

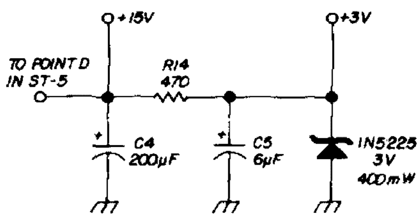


fig. 2. Power supply for the autostart, antispace circuit is connected to the ST-5 power supply (see fig. 3).

The obvious place to detect signal presence or absence is the output of the ST-5 meter amplifier at point A. The autostart input (using high impedances to minimize circuit loading) is connected here. With R2 properly adjusted, U1 will be off when a signal (+5 V or so) is present at point A and on when no signal (+2.5 V or so) is present at point A. If signal is present and U1 is off, Q1 is biased on, causing Q5 and Q2 to be biased off. When Q5 is off it allows the loop transistor in the ST-5 to control the printer magnets. When Q5 is on, the machine is locked in *mark* regardless of the ST-5 output state.

Now, back to Q2. When Q2 gets biased off by Q1 going on, capacitor C2 begins to charge through R6 and CR2. When the voltage across C2 reaches the trip point of U2 (about 3 seconds delay), U2 goes off, causing Q3 and then K1 to switch on, which turns the power on to the printer.

After loss of signal U1 goes on, causing Q1 to go off, which biases Q5 and Q2 on. With Q5 on the printer is locked in *mark.* When Q2 goes on capacitor C2 discharges through R7 and Q2. After a delay of about 30 seconds C2 is discharged below the U2 trip point, causing it to go on, which biases Q3 off, subsequently turning K1 and the teleprinter off. Diode CR3 merely protects Q3 from the voltage spike created in switching the inductive relay coil.

antispace operation

Now, how about antispace? Here you must sense the presence of a *space* condition, and after a short time delay, cause the machine to be set to *mark* even though the *space* signal persists. If you look at the output of the slicer in the ST-5 (point B in **fig. 1**) you will note that in *space* it is in negative saturation or at about − 12 volts. This causes transistor Q4 to go off which starts C3 charging through R11 and R12.

When the voltage across C2 reaches the trip point of U3, its output opens, causing Q5 to turn on, placing the machine in *mark*. The time delay set by the C3 charging circuit time constant (adjusted by R12) must be long enough to allow normal RTTY copy, but short enough to prevent annoying signals from causing the *machine to run open.*

Diode CR5 provides a low-impedance, fast discharge path for C3 to reset U3 immediately when a *mark* signal (+12V) reappears at point B. Diode CR4 merely protects the base-emitter junction of Q4 from breakdown due to the presence of the −12 volts at point B in the *space* signal condition. Diodes CR1 and CR6 allow the outputs of the autostart and antispace circuitry to be *ORed* into Q5 so that either can control Q5 without affecting the other circuits' output state. Again, since R10 is high, loading of the ST-5 circuitry is negligible.

construction

As far as construction goes, just about anything will do. I used a 3- by 5-inch perforated board with stake terminals and had room to spare. An etched board or any other *construction* technique is satisfactory. Circuit layout is not critical. Beware of mistakes in connections to Q5 since its base connections are different than you might expect.

After you have constructed your *SAA,*

some circuit setup will be required. The autostart input potentiometer should be set to produce about 1.5 to 1.6 volts at pin 2 of U1 with a *mark* or *space* signal peaked in the ST-5. Check to make sure that the circuit is operating by observing the voltage at the collector of Q1, with the antispace circuit disabled (S2 closed).

should be set so that no spikes appear at the output when receiving a normal teletype signal. However, if you don't own an oscilloscope, merely set the delay long enough to get good printout. The adjustment is not critical once you've allowed enough time in the delay for normal teletype.

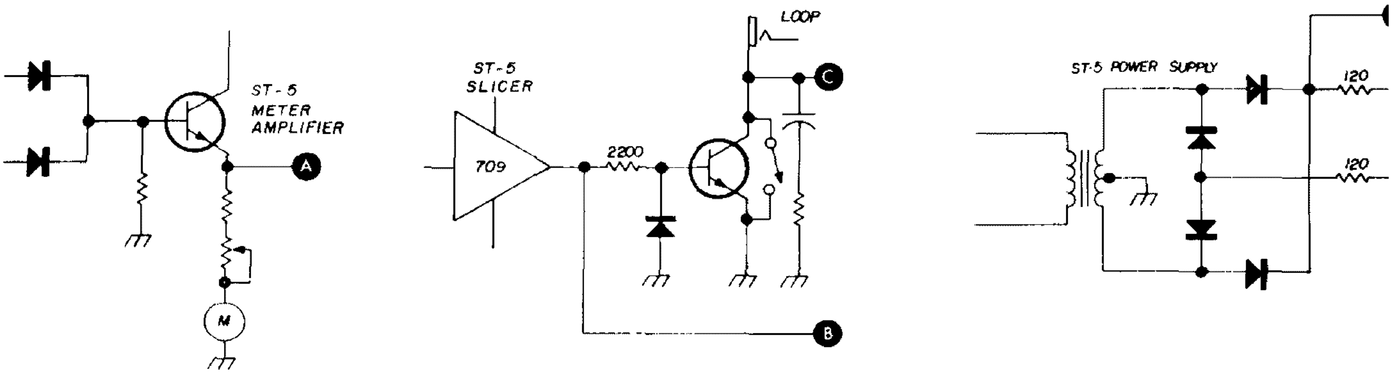


fig. 3. Connection points for the autostart, antispace circuit in the original ST-5 demodulator circuit.

As a *space* is tuned into the ST-5 space filter the machine loop should remain in *mark* until the signal goes above the trip point of U1. When this occurs, Q1 should switch off (about 1.5 to 2.0 V at its collector), K1 should then close after about 3 seconds of delay, and the machine should start and run open.

You will note that there is a slight hysteresis in the autostart circuit – the trip point to turn on is higher than the drop-out trip point. This, however, is no disadvantage since it prevents noise from initially triggering the autostart, but, once a signal trips it, the hysteresis acts to keep the printer functioning during signal fading.

setup

Setting up the antispace requires that the 100k potentiometer R12 be adjusted to provide a delay that will allow good copy, but is short enough to prevent an extended *space* condition from causing the machine to run open. This control is best adjusted by using an oscilloscope to look at the output of U3. The time delay

Switches S1 and S2 were added to allow the operator to disable the autostart and antispace functions if desired. The base of Q2 is also switched to ground during transmit to keep the autostart from turning the teleprinter off while transmitting. A most embarassing situation!

Don't try to operate the circuit from the zener-regulated, ST-5 supply. The additional loading will cause its output voltage to drop to an unacceptable level. Use the power supply connections shown in the diagram and you'll have no trouble. The ST-5 power transformer will easily handle the additional load.

That's the SAA. It has certainly made my RTTY operation less frantic and more enjoyable. Your new ST-5 and SAA may not equal an ST-6 but it comes close . . . say, an ST-5.8?

reference

1. Irv Hoff, W6FFC, "Mainline ST-5 RTTY Demodulator," *ham radio*, September, 1970, page 14.

ham radio

a single-element DX antenna

Almost unknown, the half-wave vertical can out-perform the popular ground-plane and quarter-wave verticals

B.H. Brunemeier, VQ9N, c/o 944 East Cardinal Drive, Sunnyvale, California

I have spent many years in Asia. In the Asian context, DX usually means 20- to 100-watts input on CW. A rotary beam is almost a curiosity on the CW bands as the overwhelming majority of CW DX chasers here are using a single-element wire dipole or a simple ground-plane vertical. One element, properly erected and matched, however, can produce some astonishing results.

The quarter-wave vertical, or ground plane, is too well known to require an exhaustive description. It is traditionally accepted as a very simple, and yet effective, DX antenna. However, it does have some disadvantages that are worth considering. The greatest disadvantage is its characteristic inefficiency. It is fed at a low-impedance point with a relatively high rf current. For every ampere flowing in the vertical portion producing useful radiation, there is also an ampere flowing in the ground screen. This ground-current ampere produces no useful radiation, but does account for some very significant power losses.

Most amateurs using this antenna content themselves with a ground screen of four wires, little realizing how much of

their rf power is simply warming the wires and contributing nothing to the outgoing signal. The same disadvantage applies to receiving as inefficiency in the ground system saps the incoming signal to the same degree. Yet another disadvantage is that the quarter-wave antenna just isn't very tall and doesn't have nearly the receiving capture area of a full dipole which is twice as long.

half-wave advantages

From my personal observation on the air, I've noted that the full half-wave vertical is unknown around the world. I have never yet contacted another station using one. This is indeed a mystery. The half-wave vertical has several distinct advantages which make it much more attractive than the quarter-wave. Because it is a full resonant half wave, and twice as tall, it is that much better for receiving. Its base impedance is much higher than the quarter wave, and this contributes to high efficiency. A simple example will clarify this point.

Feeding 100 watts of rf into a quarter-wave vertical with a nominal base impedance of 50 ohms would produce a current of 1.4 rf amperes. A full half-wave vertical made of typical tubing would have a nominal base impedance of 900 ohms.

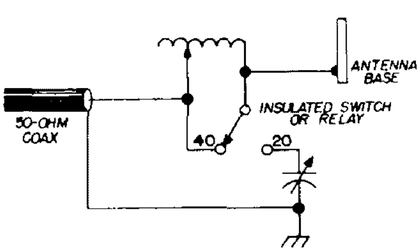


fig. 1. Switching method to use a 34-foot vertical as a half wave on 20 meters and a quarter wave on 40 meters.

Feeding 100 watts of rf into this impedance would produce a current of 0.33 rf amperes. Because the current flowing into the ground screen is the same as that which flows into the antenna, the quarter wave system would have 4.25 times more ground current than the half-wave system. The losses in the ground screen are the product of I^2R (where I is rf current and R is ground losses), and assuming the same ground screen for both antennas, the power

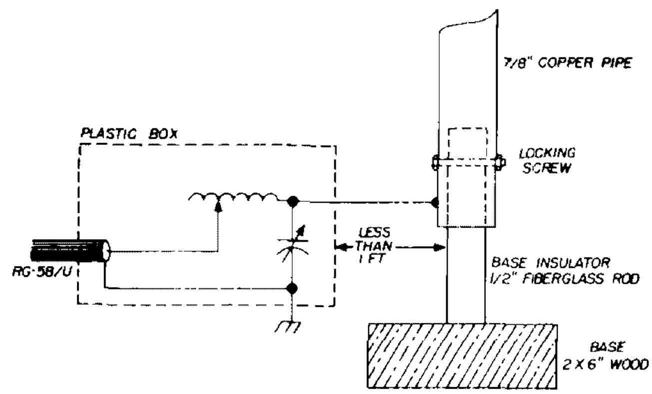


fig. 2. The antenna matching unit. The coil is 10 turns, no. 10 AWG wire, 1½-in. diameter and 1½-in. long. The capacitor is 100-pF maximum.

losses in the quarter-wave system would be eighteen times higher than in the half-wave system!

Another advantage to the half-wave system is that it has a theoretical gain of about 2 dB over the quarter wave, and concentrates that gain at a slightly lower angle above the horizon. With all these advantages to recommend the half-wave vertical, I can't help wondering why DXers around the world aren't using it. Is the 900-ohm base impedance the problem? It need not be. A simple coil and capacitor matching network takes care of that quite easily.

construction

Fig. 3 shows a half-wave vertical now in use at VQ9N. The material used is copper tubing, 7/8-inch outside diameter. It is a standard plumber's stock item on this island. Aluminum tubing is unavailable here. Note that the length is only 31 feet, rather than 34 feet, which would be a resonant half wave for 20-meter operation.

The reason for this shortage was purely economic. I bought one new 20-foot

length. It was so expensive I didn't feel like buying another whole length to cut up. A scrap 11-foot length of 5/8-inch diameter was on hand, so I spliced the two to create a 31-foot vertical. The logic was that 0.45 wavelength is so close to full resonance, that it would give essentially the same performance. This logic has proven valid in practice. Also, the supporting insulators contribute some capacitive loading, which would tend to make the antenna a little taller electrically.

The most difficult part of the project was erecting the vertical. Copper is a very soft metal and cannot support its own weight in such a length, let alone the weight of guys and insulators. During my first three attempts at erecting it, my copper column suddenly became a folded dipole in the middle. This wasn't quite what I had in mind!

On the fourth attempt I enlisted a few extra helpers. Two pulled on the upper guys, one walked up under it and the fourth pushed at the top with a long wooden pushing prop. The fourth attempt was successful, though the copper column sustained some permanent standing waves along its length, created by the earlier collapses. A vertical made of 1-inch galvanized water pipe would be much easier to set up than the copper tubing I used.

The base matching coil is made of number 10 AWG copper wire (see fig. 2). It is wound on a form and then slipped off to make an airwound coil with a 1½-inch diameter. The original coil was made with 15 turns, close spaced. The finished coil should be spread just enough so that adjacent turns don't short together. The matching capacitor is an APC air padder, 100-pF maximum. This plate spacing is adequate for rf powers up to 200 watts, which would put about 600 peak volts across the capacitor.

tuning

The matching process is a simple matter of trial and error that can be accomplished in minutes. Insert a reflected power meter or swr bridge in the line at the transmitter end and apply enough power to give some meter deflection. Begin with the full coil in the circuit, and turn the capacitor through its range. If no dipping trend is noted on the meter, remove one coil turn and repeat the process.

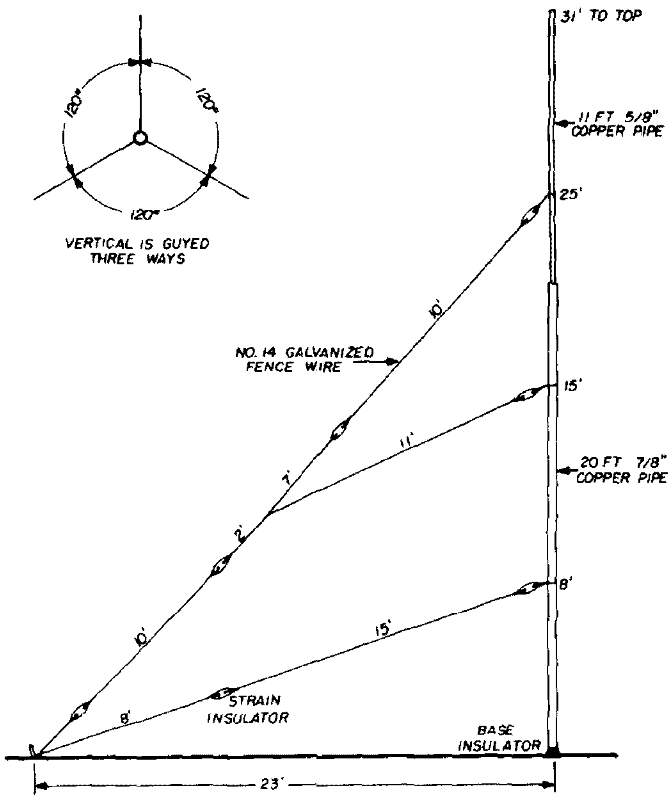


fig. 3. Overview of the 20-meter half-wave vertical. Although copper is used here, many different types of tubing could be used for the radiator.

Because the matching is quite critical, you won't see much of a meter null until you reach a point about two turns from the optimum one. Then the meter starts going down fast, and on the proper turn it can be nulled right down to zero with the capacitor. That's all there is to it.

I did my matching at 14.175 MHz, and got an swr of 1:1. The antenna response is so broad that at 14.000 and 14.350 MHz it rose to only 1.05! When the matching was finished, I had ten active turns in the circuit, which gave a coil length of 1½ inches. The unused turns were then snipped off and discarded. The

capacitor was meshed to about 60-pF.

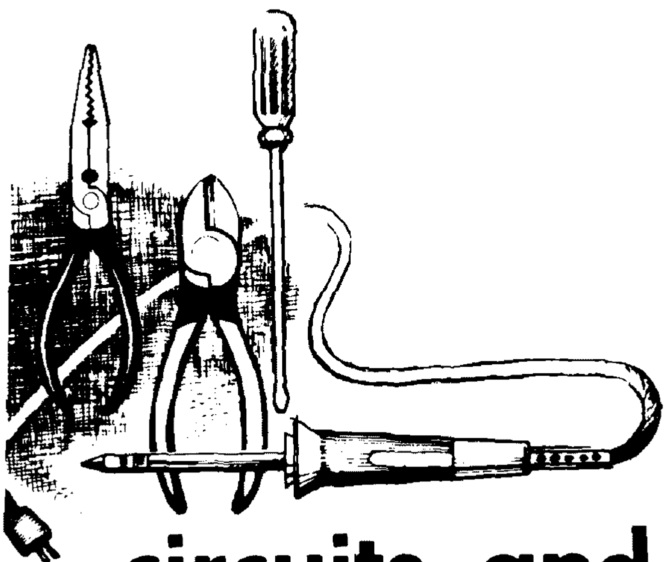
The proof of the pudding is in the signal reports. Corrugated metal roofs are almost the standard in Asia, but I went one better. My roof is corrugated aluminum, and almost level at that. A more ideal rf ground can hardly be imagined, although galvanized iron roofing does very well too. A number of tests were run on DX paths in excess of 4,000 miles to evaluate this half-wave vertical antenna in relation to other more familiar types.

I compared the half wave with the two-element quad at VQ9R and the standard quarter-wave vertical at VQ9DM (also using an almost-level aluminum roof for a ground plane). Allowing for the difficulty of taking accurate signal readings over a long path with fading, seasoned operators at the other end of the circuit gave the quad about a 6-dB advantage over the half-wave vertical.

Some of you may find it hard to believe that a single vertical element could deliver a signal only one S-unit below the popular two-element quad. Comparing the half-wave vertical to the quarter wave vertical, it was found that the half wave was considerably better. In the case at hand, the aluminum roofing rf ground plane was practically lossless for both vertical antennas.

Finally, the half-wave vertical was compared to a regular half-wave horizontal wire dipole at about the same elevation. The vertical beat the horizontal dipole by a considerable margin in any direction. So then, low-budget DXers of the world, take heart! Now's the time to pull down those wire dipoles and start standing half waves on end. At VQ9N, I run only 35 watts input on CW, and I work the world with this antenna. Where a level metal roof is not available, a ground plane of wires can come close to the same performance. An increase of signal performance over a wire dipole is very effective. Can any one imagine a simpler way to achieve so much DX gain for so little investment?

ham radio



circuits and techniques

ed noll, W3FQJ

antenna tuners

Radio amateurs in the United States and, in other parts of the world too, have developed a 50 to 70 ohm transmitter and 50 to 70 ohm antenna syndrome; manufacturers of amateur equipment have contracted the same ailment. Antenna experimenters, who probably dominate ham experimentation today, have complained for years about the lack of versatility in the output system of modern amateur transmitters. Their contention is, especially at exciter power levels, that a variety of output impedances could be made available economically and with little additional space at *least up to* 600 or 800 ohms. In fact, the transmitter with a little more versatility and a built-in swr meter and tuner might well become a very popular model.

The usual antenna experimenter prefers to work at low-power level because it is easier to obtain more conclusive results. This is not a factor that should preclude installation of more versatility in the output systems of high-powered transmitters as well, although it is true that cost and space factors are more significant for high-powered output systems.

enter the tuner

The antenna tuner unit (atu) provides the matching capability that the transmitter lacks, **fig. 1.** Its principal duty is to see that the transmitter output is matched regardless of the impedance conditions at the transmission line input. If the transmitter is made to see a proper load, it operates in an efficient and normal manner. A second fine advantage of most tuners is that they block harmonics and other spurious signals from the antenna system. This advantage holds up even when using an antenna system that can be matched directly to the transmitter.

It also is important to know what an antenna tuner does not do. It does not alter the standing wave ratio (swr), reduce attenuation or otherwise improve operating conditions on the transmission line connected between the tuner and the antenna proper. It does not improve the operating performance of the antenna. What it does do is permit you to match the transmitter to an antenna system regardless of the impedance conditions reflected to the input side of the transmission line and the other variables, serious or not so serious, that may be inherent in the antenna arrangement. An example demonstrates the above statements.

Let us assume a peak performance, narrow-band antenna is designed to operate over 100 kHz of the 40-meter band. This antenna system has been designed to permit a direct match to the transmitter output. Except for the reduction of spurious frequencies, an atu would be of little benefit in operating over this 100-kHz span.

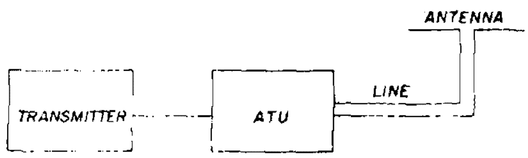


fig. 1. The antenna tuning unit on the transmitter side of the transmission line.

Off of the antenna system bandpass, impedance conditions would become unfavorable and a safe direct match would no longer be possible. Now the insertion of an atu permits the matching of the antenna system to the transmitter. Although antenna performance and line conditions are not improved by the presence of the atu, at least the antenna system can be loaded by the transmitter. In many situations the performance of the antenna system would not be noticeably different in practical communications than if the antenna were recut to this new operating resonant frequency.

Usually the impedance conditions of the above antenna resonated on 40 meters become intolerable for direct matching on 20 meters. Here again an atu of suitable design would permit you to match the transmitter to the antenna system. A least you would be able to load the antenna on 20 even though the presence of the atu does not improve the line or antenna performance on that band.

An atu is a marvel in an emergency situation and when multi-band operation is desired in a location where only a single antenna can be strung. You can at least load up the hunk of wire to obtain mediocre to good performance on a number of bands.

In summary, the atu:

Provides proper transmitter loading.

Provides harmonic and spurious signal rejection.

Permits you to accommodate an antenna that has a resonant impedance other than 50-70 ohms.

Permits you to accommodate the impedance of a broad-band, non-resonant antenna when its impedance is other than 50-70 ohms.

Permits you to load an antenna off of its resonant frequency on a given band.

Permits you to load an antenna on a band for which the antenna was not designed.

Does not change line conditions and swr.

Does not change antenna performance.

line considerations

Line factors are a consideration when using an atu at high power level. The atu does not change the line attenuation, and line attenuation does increase with the standing-wave ratio. If the line is especially long, the swr high and the attenuation per foot high, you may lose considerable power on the line even though the transmitter is matched properly.

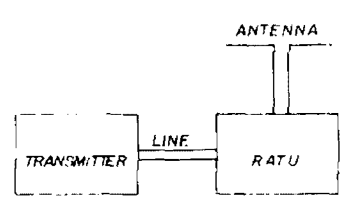


fig. 2. The remote antenna tuning unit between the antenna and the transmission line.

The power handling capability of the line is important. A high swr means voltage loops become very high on modulation crests. The rating of the line must be such that it will not break down on peaks. The higher the operating frequency the more important becomes the loss

consideration because of the increase in line attenuation with frequency.

ratu

An antenna tuning unit can be made to accommodate the transmission line as a part of a matched system by locating the tuning unit between the antenna and the far end of the transmission line, fig. 2. Minimum line attenuation and the lowest possible standing-wave ratio are now obtainable, provided the impedance of the transmission line matches the impedance of the transmitter. The remote antenna tuning unit (ratu) now matches the far end of the transmission line (same impedance as transmitter) to the antenna. This is the technique used by broadcast and other commercial transmitters that operate at a high power level.

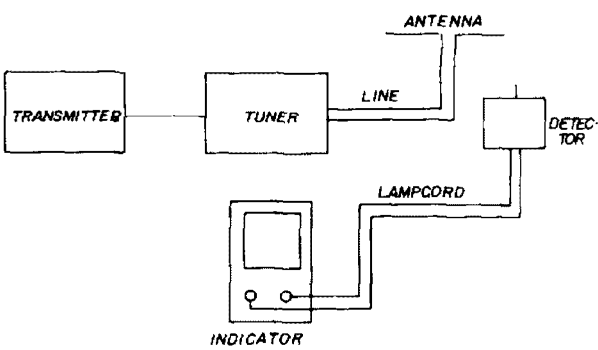


fig. 3. A remote indicator helps in adjusting the antenna tuner.

Seldom necessary in amateur radio applications, it is employed to best advantage only when the transmission line is exceptionally long and high power is to be handled. However, it is sometimes a convenient way of matching the very low resistance and high reactance of a short 160-meter antenna. When you do wish to reduce line loss to a minimum and your transmission line is not the best in terms of minimum attenuation this arrangement is worthy of consideration.

how to tune a tuner

A critical transmitter can be damaged by reflecting an improper load from the tuner. Initial adjustments must be made at low power.

Tuners come equipped with various means of band setting – plug-in coils, switched coils or switched capacitors. Regardless of the method, set the atu to the proper operating band. In adjusting a tuner try to maintain as low an swr reading as possible with the transmitter operating at low power. Usually you will have to jocky back and forth between the tuning and matching controls of the tuner to find the very least swr. In almost all practical applications this is all that is necessary in finding a true setting and minimum swr. As the power level is increased touch-up adjustments are usually necessary. Keep records of proper settings so you can return to them after changing frequencies. In most situations it is as simple as that.

false loading

False match points are found occasionally especially when using home-built tuners or trying to accommodate wide impedance differentials between input and output. Under a false condition the component values within the tuner plus the impedance conditions presented by the output load are such that most of the power is absorbed by the tuner itself. What looks to be a favorable match is reflected to the transmitter. False matches can be avoided with the use of some sort of simple field-strength indicator. The pickup should be placed as near to the antenna as is possible, fig. 3. An

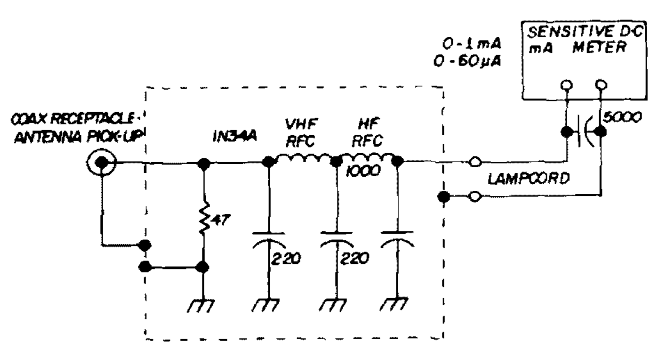


fig. 4. Simple circuit for a remote reading antenna performance indicator.

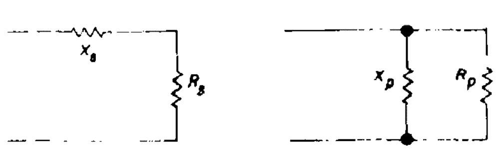
occasional check of the meter reading, using binoculars or by a second person on the job, would be appropriate if you suspect a false match point.

I have found the simple arrangement of fig. 4. helpful. A simple diode detector and output filter are used and the antenna can be a loaded 6-meter, 10-meter or CB quarterwave vertical. Suitable readings can be obtained over the entire hf and vhf-uhf bands.

A sensitive dc meter can be used as the indicator but it need not be a part of the detector proper. A long length of ordinary lamp cord can be run between the detector output and the meter. This permits you to place the meter at a point where it can be seen as you adjust the tuner. In fact, if you keep the line well filtered and isolated as far as possible from the transmission line, you can bring the meter right into the shack or at least to a point where you can see it as you look out the shack window. Proper tuner adjustment is indicated by minimum swr and maximum field reading. A false match will result in a very weak field reading.

how does an atu function?

The atu performs two major tasks. It cancels out the reactance of the antenna system and provides the resistive step-up



$$R_s \pm jX_s = \pm \frac{jX_pR_p}{X_pR_p}$$

fig. 5. Equivalent series and parallel combinations.

or step-down needed to match the resistive components. It accomplishes this by utilizing a basic characteristic of a simple or complex CL network. A series network with a specific resistance and reactance also has an equivalent parallel value of shunt resistance and shunt reactance, fig. 5. Conversely, a parallel combination also has an equivalent series value of resistance and reactance.

That such a relationship exists can be proven by setting down the expression for an equivalent parallel network of resistance and reactance as follows:

$$Zp = \frac{-jXpRp}{Rp - jXp}$$

This equation can be reworked to obtain the expressions for its real and reactive components (resistance and reactance) as follows:

$$Zp = \frac{Xp^2 Rp}{Rp^2 + Xp^2} - j\frac{Rp^2 Xp}{Rp^2 + Xp^2}$$

Note that the above is a simple series expression (R - jX). This is a fundamental series equivalent with the following values:

$$Rs = \frac{Xp^2 Rp}{Rp^2 + Xp^2}$$

$$Xs = \frac{Rp^2 Xp}{Rp^2 + Xp^2}$$

Further mathematical procedures can be used to set up the parallel reactance and parallel resistance equivalents of a series circuit. These are:

$$Rp = \frac{Rs^2 + Xs^2}{Rs}$$

$$Xp = \frac{Xs^2 + Rs^2}{Xs}$$

In matching an antenna system to a transmitter or line, an appropriate network (atu) is inserted between the series resistance and reactance presented by the antenna to reflect an equivalent parallel impedance that matches the strictly resistive impedance of the line or transmitter. (Sometimes the load too has a reactive component that must be considered.) The *parallel-connected network* of fig. 6 consisting of a series inductor and parallel capacitor can serve as a

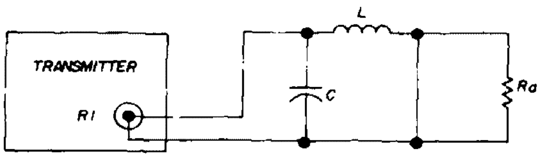


fig. 6. Simple LC matching arrangement.

simple matcher. The values of these components are determined and become of such value that an appropriate impedance match is made between the two resistive components R1/Ra. Let us assign a symbol of "n" to the latter ratio.

Further mathematical procedures can now be used to reduce the equivalent series and equivalent *parallel reactance* equations to the following simple expressions:

$$Xs = Ra\sqrt{(n\text{-}1)}$$

$$Xp = \sqrt{\frac{nRa}{(n\text{-}1)}}$$

Let us assume we are to match a 72-ohm transmitter to an antenna with an impedance of 36 ohms resistive and -160 ohms reactive (capacitive), fig. 7. To *balance out* the reactive component of the antenna it will be necessary to use a series inductor with at least an inductive reactance of +160 ohms. Additional inductive reactance will be necessary to handle the impedance match. Likewise the reactance of the shunt capacitor must be selected for appropriate impedance match. The two equations are now employed. Additional series inductive reactance needed is:

$$+Xs = 36\sqrt{2 - 1} = 36 \text{ ohms}$$

$$-Xp = \sqrt{\frac{2 \cdot 36}{2 \cdot 1}} = 72 \text{ ohms}$$

The former is added to the previous 160 ohm value to obtain a required L value of:

L1 reactance = 160 + 36 - 196 ohms

The parallel reactance value becomes:

C1 reactance = 72 ohms

The above reactances can be converted to inductance and capacitance at the operating frequency by using the basic reactance equations:

$$L = \frac{Xl}{2\pi f} \qquad C = \frac{1}{2\pi f Xc}$$

If operation is centered about 1.82 MHz, actual values are as follows:

$$L = \frac{(196)\ 10^{-6}}{(6.28)\ (1.82)} = 17.1\ \mu H$$

$$C = \frac{10^{-6}}{(6.28)\ (1.82)\ (72)} = 1216 \text{ pF}$$

When antenna characteristics are not known exactly it is no great problem. You can assume very approximate values or draw from your practical knowledge of coil and capacitor sizes for a specific frequency. It is then only necessary to make one or both of the reactances adjustable. This is why antenna tuning units are indeed tunable. They permit you to adjust the matching network for an idealized match on any frequency by making adjustments and watching an swr bridge for the very best match and forward output.

counters

Thank you, Roy (R.W. Lewallen, WØETU), for sending the helpful counter data that follows:

In reference to your article in the July issue of *ham radio,* I would like to contribute a systematic method of wiring a divide-by-n counter, which besides not

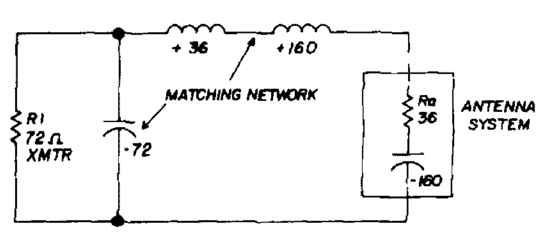


fig. 7. A typical antenna matching situation as explained in the text.

All of the above is quite understandable. However, to the uninitiated, the wiring of an integrated circuit appears to be a very complicated thing. Actually it is simple and the major complication is usually the printed-circuit board. However, this can be avoided by using straight wiring techniques as suggested in the first experimental procedures in the June column. If you use binding posts and jumpers it is also possible to change the count sequences between the combinations shown in figs. 2 and 3.

The pin-out wiring diagrams for the 7490 are given in figs. 4 and 5. Note how very simple it is. There are a number of terminals to which no connection is made and another group which are all tied to common. Of course, there are supply voltage as well as input and output connections to be made. The diagrams of fig. 4 are for using the 2-to-1 and 5-to-1 counters separately. Both the connections of fig. 5 provide the 10-to-1 count.

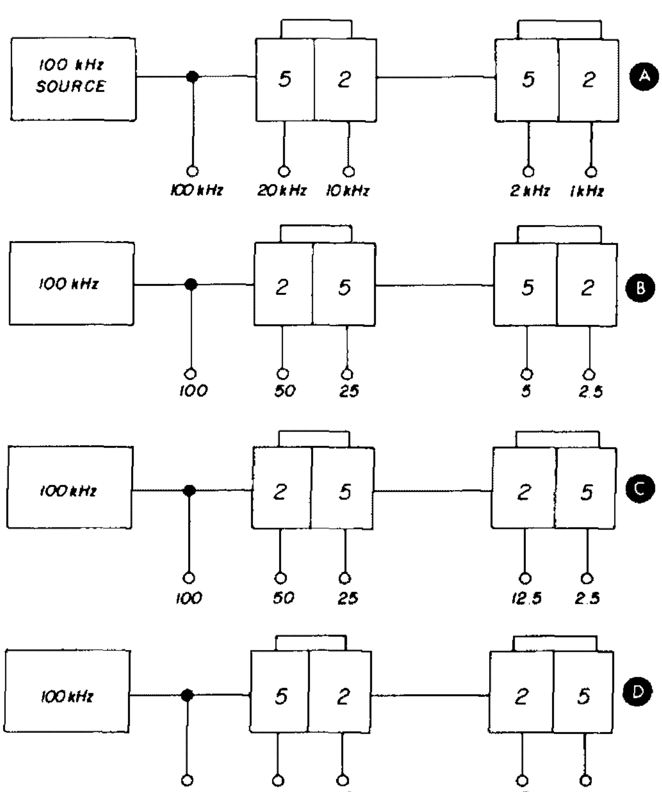


fig. 2. Some of the count possibilities using two decade dividers.

However, in the first example, the first count is 5-to-1; the second, 2-to-1. The second example is the converse, using the initial 2-to-1 count and then the 5-to-1.

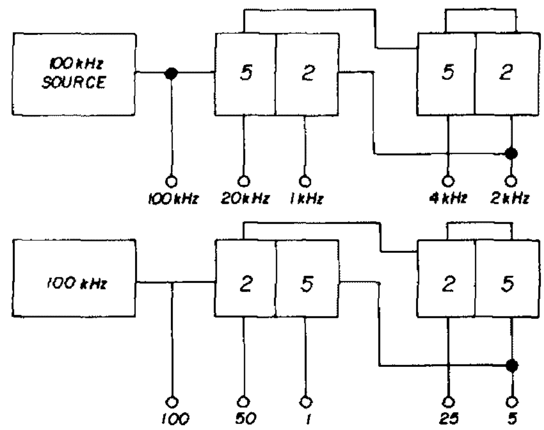


fig. 3. Two additional count possibilities using two decade dividers.

digital IC oscillators

Digital ICs of suitable design can also be used as high-frequency crystal-controlled square-wave generators. The 7400 NAND gate used initially in this series can be operated as a high-frequency oscillator. Two of the four gates are wired as a multivibrator while a third one is used as a buffer output. Doug Blakeslee, W1KLK, has used this common IC successfully with the circuit of fig. 6A.[1] Its output is followed by two 7490 decade dividers.

Ted Bensinger, W5PCX, uses the 7400 in the 3-MHz IC oscillator arrangement of fig. 6B.[2] Two of the NAND gates again serve as the multivibrator while the two other sections are pressed into service as buffer and calibrate outputs. Two decade dividers provide the countdown to 30 kHz. W1KLK operates his circuit at 3 MHz to get the same 30-kHz output. However, he employs a high-frequency 74H00 NAND gate. Theoretically this IC should provide steeper sides and higher harmonic output levels.

the ham notebook

oil-filled capacitors

When is comes to dependability, you can't beat oil-filled filter capacitors. While electrolytic capacitors have the advantages of compactness and low cost, they just don't have long lives. Whenever long-term stability and dependability are important, design engineers invariably specify oil type filter capacitors. More amateurs would be inclined to use them but for the fact that they are reputed to be far beyond the amateur's price range. This is not so in all cases, for surplus oil-filled capacitors appear on the market at quite reasonable prices. Some of these first-rate capacitors are rated in "working ac volts." Lacking a direct translation for this industrial rating, the average amateur will often write off these bargains.

For oil-filled capacitors, commonly used with ac, there is a unilateral conversion table for their utilization with dc voltages (**table 1**). For various reasons, there is no equivalent conversion setup for dc-to-ac, but this is of no interest for the amateur constructor.

An alternate arrangement, which has served me very well over the years, is as follows: Multiply the ac voltage listed by 2.828. What this equation shows is that the maximum steady dc voltage is equal to the peak-to-peak *ac voltage rating*. This calculation will give you the *maximum* dc voltage rating, and for the sake of conservative engineering and trouble-free operation on rectified 60-Hz ac, it is wise to de-rate the dc maximum voltages given by roughly one third.

The calculations listed above seem to work best for the higher ac voltages, and it roughly parallels the equivalent voltages in the higher ranges of table **1**. It is worth noting that you may end up with some odd voltages, such as 2121 volts or 2750 volts. Do not allow this to confuse you, since that seemingly odd value is very close to the true rating.

In some cases, if the actual size of the oil-filled condenser in question is known, it may be possible to identify its equivalent maximum dc voltage rating by comparing its size to a dc capacitor which is catalogued and rated by the manufacturer. Armed with the foregoing knowledge, it is possible to match up the various offerings which appear from time to time. Still in doubt? Recently I picked up a 13-μF oil-filled capacitor, rated 950 Vac, equivalent dc rating approximately 2700 volts. This is the maximum rating, and when used at roughly 2000 volts, it should last a lifetime. The cost, utilizing the above information, was only five dollars.

Neil Johnson, W2OLU

table 1. Dc working voltages of ac rated oil-filled capacitors. All dc voltages listed are the nearest standard voltage.

ac working voltage	dc working voltage
110	200
220	400
330	600
440	1000
550	1500
660	2000

NE561 as an ssb detector

I presently have on the drawing board a receiver which will use a 561B as a multimode detector (fm, a-m and ssb). Needless to say, I was somewhat dismayed to read in the March 1972 "Circuits and Techniques" column that the circuit will not work as an ssb detector. I decided it was time to do some breadboarding.

In the September 1971 issue of *ham radio* WA2IKL was close. A block diagram of his detector in the sideband mode is shown in fig. 1. The crystal oscillator locks up the PLL and the output from the vco is fed to the balanced modulator.

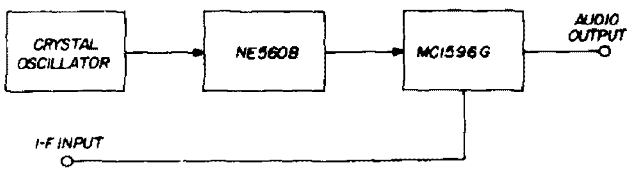


fig. 1. WA2IKL's detector circuit in ssb mode.

The NE561 is in essence an NE560 and an MC1596 combined into one package. I had assumed that the 561 would work perfectly as a multimode detector. I was given further encouragement by the Signetics applications memo on the PLL which stated, " . . . Its design is similar to the Signetics 560 Phase Locked Loop but it contains an additional product detector to perform the a-m detection function."

The block diagram of a 561 operated in the a-m mode is shown in fig. 2. Phase detector number one serves only to lock up the vco to the a-m carrier. This detector makes no use whatever of the modulation sidebands. The vco is locked ninety degrees out of phase with the a-m carrier; therefore, when an external ninety-degree phase shift network is used the vco will be in phase with the carrier. The a-m detection occurs in phase detector number two. In the fifties, sidebanders used to call this exalted carrier detection, except for the fact that they did not sync their bfos to the incoming signal.

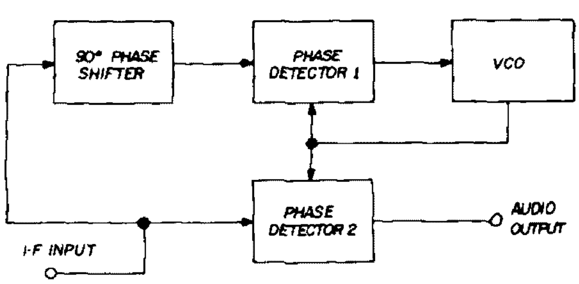


fig. 2. Phase lock loop used as a-m detector.

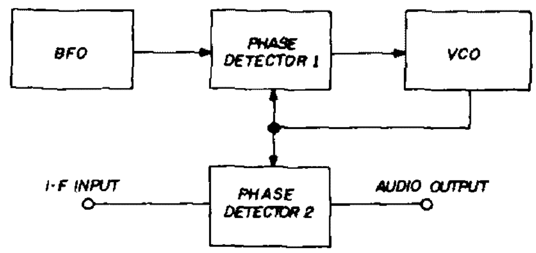


fig. 3. Phase lock loop used as ssb detector.

Because a sideband signal is transmitted without its carrier, the missing carrier must be reinserted at the receiver. There is no point in combining the bfo with the i-f signal directly. As shown in fig. 3 the bfo signal is fed to the fm input to lock up the loop. The i-f signal is fed to the a-m input to be detected in the second phase detector. The ninety-degree phase shift network would be meaningless in this case. The vco frequency is identical to that of the bfo, and the vco becomes the reinserted carrier. Detection occurs in phase detector number two just as in any other product detector.

The circuit which I breadboarded is shown in fig. 4. The 455 kHz i-f signal was stolen from the Drake 2B through the Q-multiplier socket. The audio was fed into the receiver through an audio input I had added to the 2B earlier. The third converter tube was removed from its socket to disable the unused parts of the receiver.

While there are quite a few omissions in the design, sophisticated circuitry was skipped in the name of speed. This is not intended as a construction project. The hope is that it will give some good ideas.

The performance of the detector is

excellent. Noticeable distortion both by ear and by oscilloscope sets in at about 2.5 volts peak-to-peak output. Some high-frequency bfo hiss is heard, but a good low pass filter would probably take out most of it.

My ultimate receiver will have an NE561B as a multimode detector.

Max Robinson, K4ODS

however, had only one rf stage ahead of the mixer, which means the image response was not the best. When tuning near the low end of 20 meters, for example, the image from a strong Loran station completely wiped out the lower 5 kHz of this band. Further up the band, the image from a foreign phone station dominated another segment. The receiver

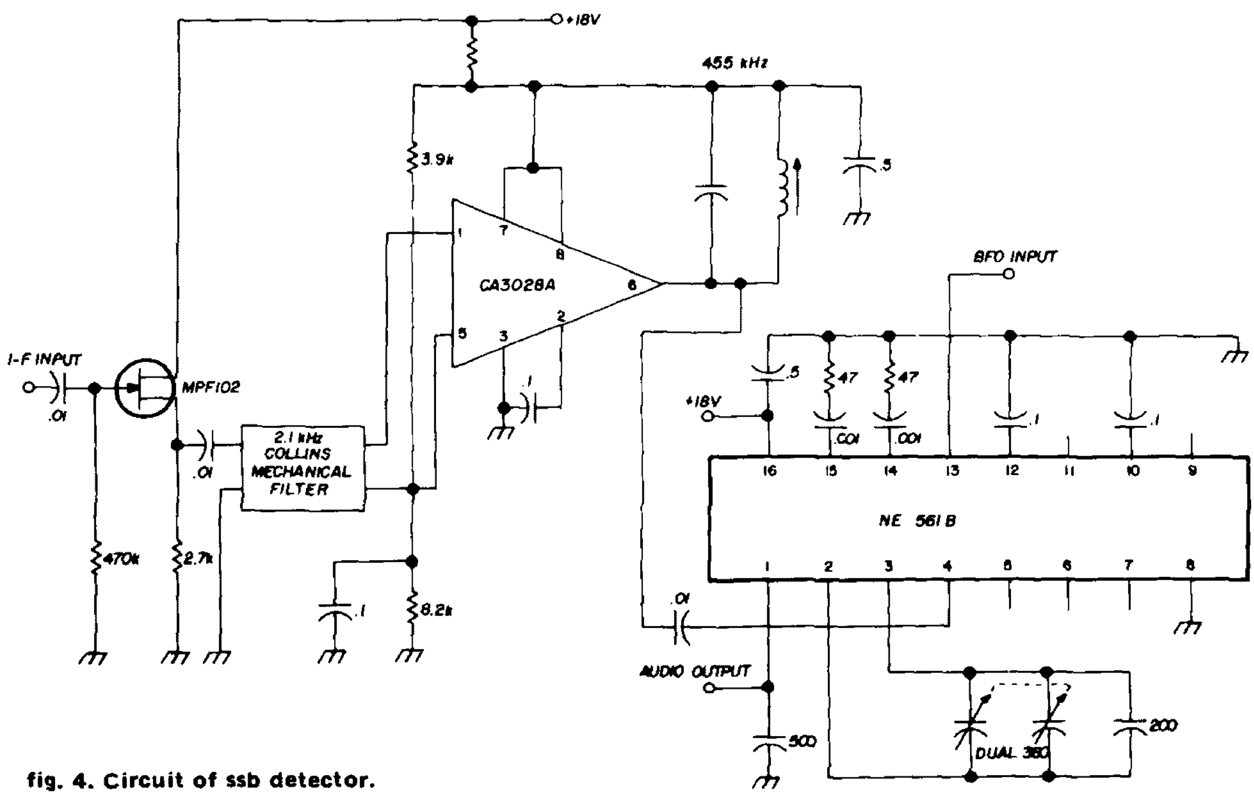


fig. 4. Circuit of ssb detector.

receiver image suppression

At a local ham swap meet I found what appeared to be a good bargain: a refurbished receiver about five years old with several interesting features, including a built-in Q multiplier. I wanted to give the set a smoke test before making the purchase, but the only source of power was being used to operate a PA system over which frustrated wives were trying to locate lost kids and husbands. Anyone who has ever been to a ham swap meet will know what I mean.

I bought the receiver anyway; it looked to be in mint condition. The set, was practically useless for chasing weak DX signals.

Images are easy to recognize, since they appear on the dead-zone side of zero beat in receivers with good i-f selectivity. The signal-to-image ratio in superhets can be improved in several ways, including the addition of more front-end selectivity, multiple conversion, and special circuits in the mixer input. However, I didn't want to dig into the set, so an alternate solution was needed.

The outboard trap shown in **fig. 5** is about the simplest means of attenuating images, without opening up the receiver. The LC circuit was built into a metal box to reduce hand capacitance, which makes

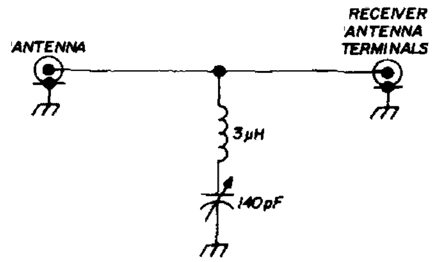


fig. 5. Series-resonant trap for improving receiver signal-to-image ratio. Circuit covers 20 and 40 meters.

tuning difficult. The trap was also effective in attenuating a strong local ssb signal that caused severe receiver overload. This device has been published in the literature many times, but I offer it again for those who may have overlooked it.

While certainly not a cure-all for receiver front-end problems, this simple circuit allowed weak cw signals to be copied that could not otherwise be heard on the 20-meter band.

Alf Wilson, W6NIF

neutralizing tip

With the tight packaging used in modern final amplifier design, it's sometimes difficult to find space for a neutralizing capacitor. The neutralizing scheme used in my mobile rig is shown in fig. 6. A one-inch strip of copper foil (shim stock) was formed around the final amplifier tube envelope and positioned so that it was level with the plate. The foil was

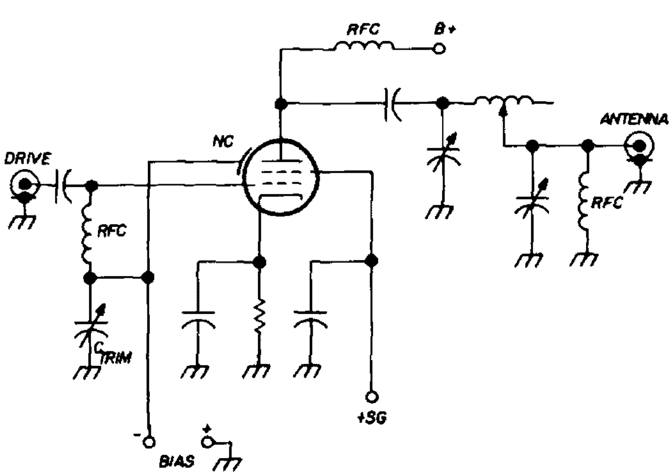


fig. 6. ZE6JP's neutralizing scheme for limited space. Neutralizing capacitor, NC, is a strip of copper foil placed around the tube envelope.

connected to a compression-type trimmer capacitor, and the amplifier was neutralized in the usual manner. A little cement on the copper prevents shifting relative to the plate.

Try the circuit first without the trimmer capacitor. It may not be needed, depending on distributed capacitance in the circuit and the interelectrode capacitance of your particular tube. Note the rf choke connected across the amplifier output. This component is sometimes omitted in pi-network amplifiers, but it's good insurance against high voltage appearing on the amtenna should the plate blocking capacitor develop a short circuit.

H. L. Booth, ZE6JP

spurious signals with the Yaesu

A number of obviously spurious ssb and CW signals have been heard from the United States and Japan recently. Investigation led to writing Yaesu for help. Yaesu's president, JA1MP, was very cooperative, and should be thanked for his assistance.

One of the signals was on lsb on 14087.37 kHz. It was caused by an usb signal on 14306. 85 (a mean frequency of 14197.11 kHz). JA1MP says that several trap coils are used in Yaesu equipment to reduce spurious radiation by at least 50 dB.

In this case the spurious signal was in an FTdx-400, and was probably caused by mistuning of the trap coils L17 and L19 which are located in the plate circuit of the transmitter first mixer. This spurious crosses at about 14,200 kHz and is strongest at that frequency. His suggestion for alignment is that the transmitter be tuned to 14,220 kHz and the receiver to about 14,180 kHz where the spurious is heard. Then adjust L17 and L19 for minimum S-meter reading on the receiver. When properly tuned, the spurious is down more than 50 dB — even at the worst point.

The CW spurious signals were heard on the 10-meter band, where they are caused

in Ftdx-560 transceivers by the second harmonic of the 3180 kHz i-f that is generated by the transmitter second mixer stage when the mixer is overdriven. Especially on 28 MHz, users are apt to overdrive the rig to overcome the lower efficiency due to the higher frequency. To reduce the second harmonic, Yaesu now is modifying all rigs to install a sharp suck-out crystal filter in the i-f circuit of transceivers.

JA1MP enclosed a copy of the Spectronics "Yaesu Information Bulletin" relating to the FTdx-560/570 equipment. It shows how to place a 6358.6 kHz crystal, XT-1, across TC-3 (the middle hole in 8PF-5) which tunes a tank circuit in V-203.

Bill Conklin, K6KA

supply, you may prefer to follow their guidelines if you want the latest in solid-state design. The April 1972 issue of *73 Magazine* had an article on a dc supply for the HR-2. This article had "current limiting," if you want to consider a resistor in series with the bridge rectifier

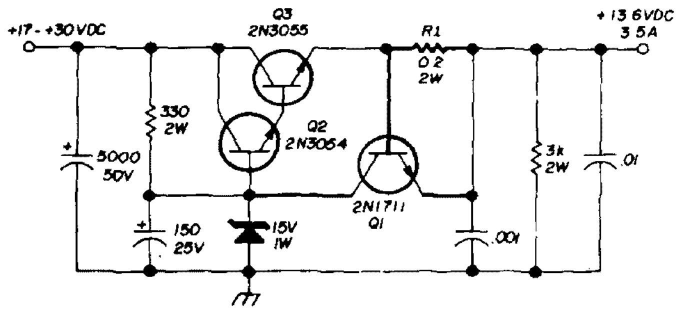


fig. 7. Adding current-limiting to an existing power supply. Q1 and R1 are new components. See text for calculating value of R1.

current limiting

The current list price for a 2N3054 is $1.20, for a 2N3055 it is around $2.00 and for a 2N3716 it is around $6.50 (all in quantities of less than ten). For the price of a 2N1711, 75c to 85c, plus the cost of a 0.2 ohm resistor, you can save yourself many dollars in replacement costs. If you have fuses in your supply, you can save on these also. Fuses are nice, but they are just not fast enough for today's solid-state devices.

If you are using a zener-regulated supply with a series pass transistor, by merely adding a series resistor and transistor switch combination, you can have whatever range of current limiting you desire.

Fig. 7 is representative of many of today's dc supplies; however, there may or may not be a Darlington pair as I have shown here. Numerous articles have been written covering supplies for the current breed of vhf transceivers; however, some are worse than others. Some say they have current limiting, but close inspection reveals the output pass transistor is not protected as all. The November 1971 issue of *QST* had an excellent article on dc supplies. If you are building a new to collector of the pass transistor any form of limiting. This method does not prevent the pass transistor from getting extremely hot and eventually wiping itself out.

Refer back to fig. 7, with the added components Q1 and R1. If the current drawn by the load, exceeds a preset value (I1R1), Q1 then conducts depriving the base of Q2 of its drive voltage. The output goes to "zero" and nothing burns out. The added transistor does not need to be mounted on a heat sink, and is a TO-5 case device. The pass transistor and its Darlington driver should be mounted on heat sinks and adequately insulated with silicon grease.

If your supply does not have a Darlington driver, then connect the collector of Q1 to the pass transistor base. Just imagine that Q2 is not there.

To calculate the value for resistor R1, use the following formula: R1=(0.7)/Isc, where Isc is the short-circuit maximum current desired.

This addition to my supply has saved many power transistors from destruction. It is a very welcome addition. Thanks to Bill Durspek, WØBVR, for his help in solving my problem.

William P. Lambing, WØLPQ

comments

quartz crystals

Dear HR:

It is commendable that amateurs are becoming aware of the intricacies of quartz crystals; it is great that two authors chose to write on the subject in *ham radio.* The subject is a clear-cut *science and not* the black magic which some people try to make it. As a result of the black magic aura, much misinformation exists. The prime movers in the field of crystal enlightenment are the US Army Electronics Command, Fort Monmouth, New Jersey, and Bell Telephone Laboratories. Annually, interested factions of the electronics industries are brought up to date at the Frequency Control Symposium held in Atlantic City, New Jersey. IEEE-sponsored committees are continuously at work on a national and international basis to aid in standardizing terms, measurement techniques, and manufacturing procedures so that everyone can communicate on a common basis.

It would behoove the amateur radio operator to learn the terms used in the industry as he studies crystals. Author Sondgeroth interchanges terms that have specific definitions other than the meaning he intends. Such will eventually confuse any student if it has not already. While I personally have differed with the parlance of the crystal industry, I find it is necessary to use the language understood by my converser.

The term "holder" refers to the case in which the quartz is mounted. The holder is only a minor contributor to C_o. Static capacitance (C_o) is primarily a function of the electrodes sandwiching the quartz. Why must the author refer to C_o as holder capacitance? Depending on the area of the plating, and the thickness of the quartz, C_o can range from 3.5 to 7.0 pF on crystals commonly found in surplus houses. The figure of 5 pF is only a seat-of-the-pants estimate.

Elements of the motional arm are usually designated C1, L1, and R1, although C_M, L_M, and R_M are understandable. When the subscript M is used, I can't see the reasoning behind defining the terms as "equivalent" and "effective." I believe this is misleading if not incorrect. "Equivalent Series Resistance" is the term used to define the resistance of the crystal in an oscillating circuit and it includes series resistances and parallel conductances of the holder, not shown in this equivalent circuit. Hafner* uses the term equivalent a second time as equivalent reactance (X_e) of an oscillating crystal, but this cannot be broken down into the reactances of C_M and L_M,solely. The reactance of C_o also has an effect. The terms motional capacitance, motional inductance and motional resistance are preferred.

A point on which the reader should also be cautioned is Mr. Sondgeroth's *statement* that you can *save money* by tuning a lower accuracy crystal. Not necessarily so! Lower accuracy crystals are generally so designated because they have poorer temperature characteristics. If you operate under conditions of wide temperature variations, you may become a slave to the technique of crystal tuning.

It is not my intent to sharpshoot the article — I feel it is well written and pertinent to an amateur's problems.

*Standard Definitions and Methods of Measurement for Prezoelectric Vibrators, IEEE No. 177, May, 1966.

In G8ABR's contribution to the *ham notebook,* a statement is made that air-mounted crystals, e.g. 10X, will always oscillate at exact multiples of the fundamental. I believe in this case the crystal is vibrating in the fundamental mode and tripling electrically. Most of the air mounts (pressure types) will not operate in harmonic mode because the coupling of the plates to the quartz is poor. This would include FT-243 and older HC-6/U crystals. It is just about impossible to make any crystal with exact frequency multiplication by operating in harmonic modes.

Don Nelson, WB2EGZ
Voorhees, New Jersey

code practice

Dear HR:

The editorial in the May 1972 issue concerning building up code speed with coastal station transmission copy practice is certainly meritorious. Table 1 is part of the schedule I compiled and distributed to the fleet of off shore tankers I am connected with, for use by the Radio Officers. Data was secured from various publications and direct observations and checks.

table 1. Weather transmissions in International Morse Code.

station location	time GMT	frequency kilohertz
WCC	0050	436 2036 4331
Chatham,	1250	436 2036 6376 8630
Mass.	1650	436 2036 6376 8630
WSL	0500	418 8514
Amagansett,	1100	418 8514 13024.5
New York	1700	418 8514 13024.5 17021.6
	2300	418 8514 13024.5 17021.6
WSC	1418	460
Tuckerton,	2318	460
New Jersey		
WMH	0130	428 8686
Baltimore,	1600	
Maryland	1330	428 8686 12952.9
	1930	
WOE	0105	472 6411.35 8486 12970.5
Lantana,	1605	
Florida		
WAX	0135	488 4295 8526 13011
Miami,	1335	488 8526 13011 17199.2
Florida		
WPD	1800	420 13051.5
Tampa,		
Florida		
WLO	1300	438 8714 12704.5
Mobile,	1700	
Alabama	2300	
WNU	0430	478 4310 6495 8570
New Orleans,	1630	478 6495 8570 12826.5
Louisiana		
WPA	1748	416 8550 12839.5
Port Arthur,		
Texas		

WSL transmits about a half hour weather copy four times a day as indicated. It is tape and about 23 or so wpm. WAX is good for two daily periods at about 16 wpm, also tape. WOE sends two daily periods, about 20 or so wpm, and most of the times is good copy. WSL on the 12-MHz frequency seems to be about the best bet for the most powerful signal.

About the only maritime telegraph stations I have found that send in the 30- to 35-wpm category are the Russians. Their merchant fleet, in recent years, has expanded phenomenally. They met the need for new frequencies without trouble. They picked out low-powered coastal stations and moved on top of them. They stayed clear of the high powered ones like WSL or WCC. One example is UQK, Riga, who is parked on WLO, Mobile, Alabama, 12704.5 kHz. They are netting in as the ships use the same frequency as the coastal stations. Weather transmissions from their ships to UQK is at about 0000, 0600, 1200 and 1800 GMT. Speed is in the 30- to 35-wpm range, very good code sending addressed to *pagoda* and is in standard international code. These messages are mostly number groups with ships names and should be good practice for our high speed boys.

I would suggest stressing the matter of *regular daily* code practice to build up speed. Once or twice a week will be of not much help in building up speed. A month or so of daily practice should bring a 13 worder up to 20 plus with no strain.

Paul Szabo, WB4LXJ
Tampa, Florida

power in reflected waves

Dear HR:

Various letters have been received, commenting on my paper, "Power in Reflected Waves" (*ham radio,* October, 1971). Most writers agreed, but some disagreed with my principal conclusion that there is no power in reflected waves on a transmission line. I wish to thank all who wrote me, for their interest in my paper.

Much of the disagreement appears to be based on incredulity, rather than on reasoned technical analysis. This is understandable, because most pertinent writings which have appeared in amateur periodicals, and indeed in important handbooks written for amateurs and even for professionals, have discussed power in reflected waves as if it were a reality. Rare exceptions are 'Losses in Feedlines" by Byron Goodman, *QST*, December, 1956, and "The Mismatched RF Transmission Line" by Carl C. Drumeller, *73 Magazine,* November, 1969. These writers correctly said that so-called *reflected power* is not really power at all.

The technical criticisms received can be combined and summarized as follows:

1. The voltage and current standing waves on an unmatched transmission line have a phase difference of 90 degrees only when reflection is complete, and therefore, under other circumstances they represent power.

2. The input impedance of the transmission line is not matched, and cannot be matched, to the output impedance of the transmitter unless the tube's load resistance is equal to the tube's plate resistance, which is not the case in practice; therefore, virtually complete reflection of the *power* in the reflected wave does occur at the transmitter output.

3. In the case of my fig. 2, there is a reflected wave on the coaxial line shown, for the reason that there is no other possible destination for the *power* in the reflected wave on the open transmission line.

I will discuss briefly these three points in order.

Point 1. The fact is, the voltage and current standing waves on a mismatched transmission line have a phase angle of 90 degrees even though reflection is not complete. To avoid going into a detailed proof, the references are cited in evidence.

Point 2. This criticism is incorrect, but in any event it is irrelevant; it does not prove or demonstrate that there is power in the reflected wave.

Point 3. This *is* a peculiarly circular argument. It claims that I am wrong in denying that there is a reflected wave on the coaxial line in my fig. 2, because, contrary to my principal conclusion, there *is* power in the reflected wave on the open transmission line, *and this must* appear on the coaxial line. In other words, it says that I am wrong because *I* am wrong! This criticism, like the one above, does not prove or demonstrate that there is power in the reflected wave on the open line.

Whoever originally wrote about power in reflected waves on transmission lines as being a reality, no doubt thought that this concept would serve to simplify, for the non-professional, the manner of formation and significance of standing waves. Historically, it has served only to complicate the matter endlessly, as writer after writer, following the original lead, and unwilling to break with precedent, has grappled with reflection of *power* in rf lines and its re-reflection back and forth ad infinitum, with trying to explain how it is that a directional wattmeter can indicate, under some circumstances, more power in the line than the transmitter is putting into it, and how power at the same frequency can travel both ways simultaneously. They seem to have forgotten that the basic definition of power is, "the rate of doing work," and have failed to show where and how this work, corresponding to the assumed power in

the reflected wave, is being done. Of course, they cannot show this, because it is not power and *therefore* is *not doing* any work.

The underlying error in these misconceptions is the failure to distinguish between ac power, or volts ac times current ac, multiplied by a power factor other than zero, from volts ac times current ac, multiplied by zero power factor. The former is power; the latter is not, and it is a fundamental error to call it so. Power utility engineers know better than that.

The basic point to be recognized about power in a transmission line is simply that the power from the transmitter into the line is the sum of the power lost in and from the line and in any additional matching or other devices inserted into the line and the power delivered to the antenna. There is no other power moving in either direction. It is that simple.

Hubert Woods
Jalisco, Mexico

tuning toroidal inductors

Dear HR:

In his article, "Tuning Toroidal Inductors," (*ham radio*, April, 1972), author WAØJYK indicates that a grid-dip oscillator cannot be used because there is not enough flux leakage from the toroid.

The fact is that a grid-dip oscillator will give excellent readings on a tuned circuit having a toroidal inductance. Just put a loop of wire through the toroid and twist it into a link around the coil of the grid-dipper. If a precision capacitor is used, the inductance can be calculated to a degree of accuracy limited only by the care with which the resonant frequency is read.

Even rough checks by the gdo-capacitance-frequency method can give better results than those given by all but the best laboratory bridges since the measurement is usually made with the inductance excited at the frequency at which it will actually be used.

Barry Kirkwood, ZL1BN
Auckland, New Zealand

reciprocating detector

Dear HR:

I have received several letters regarding my "reciprocating detector" article which appeared in the March, 1972, issue of *ham radio*. Transistor Q5 is part of the reciprocating detector switch, but the questions are understandable due to the lack of a dot to show a connection in the schematic; resistors R4 and R5 should be joined with a dot where these two resistors form a junction point at the input to the diode and the base of Q5. The diode is a 1N252.

Several readers have also asked where the selectivity curve is 500-Hz wide and what is its slope. The filter I used was designed to have its 500-Hz passband at the 3-dB points on a slope which is not particularly steep for an inductive filter. Indeed, at 500 Hz, the L3 inductance is very loosely coupled to the other two sections of the transformer. The bandpass formula (f_r/Q_o) indicates that the bandpass of the filter is actually narrower than 500 Hz – in fact, bandpass is closer to 250 Hz. The 390-ohm resistor used in series with one of the differential inputs loads the thing down so it is broader. If the bandpass is too narrow, poor lock-in range is experienced on a-m, and there is very poor "presence" in the quality of ssb signals. If the bandpass is too wide, poor impulse rejection will result.

Stirling M. Olberg, W1SNN
Waltham, Massachusetts

RTTY speed converter

Dear HR:

I just completed construction of the RTTY speed converter described by WA6JYJ in the December, 1971, issue of *ham radio*, and it works like a charm. I built the converter on a printed-circuit board which greatly simplified its construction. I can furnish printed-circuit boards to interested readers for $6.00.

Earl E. Palmer, W7POG
17510 Military Road South
Seattle, Washington 98188

radio control

Dear HR:

I would like to remind your readers that the frequencies, 53.10, 53.20, 53.30, 53.40 and 53.50 MHz have been recognized by the FCC as *radio control frequencies* for licensed amateurs who engage in remote control of model boats or airplanes.

Interference on these radio control frequencies has been on the increase, causing loss of control. This can be disastrous to the model builder who has spent countless hours and a lot of money on his model, only to see it crash because of interference.

Considering all the frequencies available to amateurs who use six meters for communications (CW, ssb, RTTY, etc.), it seems reasonable to ask them to stay clear of the radio control frequencies noted above. In addition, since it is impossible to build fancy receivers into the very small space available in most models, it would be appreciated if a reasonable guard band, say 6 kHz, could be observed.

Pierre J. Catala
Needham, Massachusetts

pi-network inductors

Dear HR:

W6FFC's article on pi-network design is easily the best and most comprehensive treatment of the subject that has ever been published in a ham magazine. Congratulations. I believe I can add something on the matter of inductors for high-power tank circuits.

There has been a lot of theorizing on the effects of corrosion on bare copper, and the benefits of silver plating, but little actual measurement. Some time ago, out of curiousity, I resurrected an old 10-meter tank coil from my junkbox. It had been wound at least ten years previously, and consisted of several turns of 1/4-inch copper tubing, 2-inches long. It was the familiar chocolate brown color of old copper.

I measured the Q of this coil on a freshly-calibrated HP 260-A Q Meter. It measured 173. Next, I had the coil chemically cleaned and brightened. The Q increased to 176. Then the coil was silver plated .0002 inch. This raised Q to 178. All of these measurements were made during an eight-hour period.

As a final experiment, I wound a coil of number-14 tinned copper bus wire, of the same length, diameter and inductance as the copper tubing coil. The Q measured 172.

These measurements show that the benefits of silver plating are negligible at frequencies up to 30 MHz. The difference in efficiency of an amplifier using any of these coils could hardly be measured. However, I would not recommend the use of the wire coil, since, as W6FFC points out, at 30 MHz the coil dissipation may be as much as 100 watts, and the wire coil would be inadequate. Dissipation rather than Q is the real reason for using tubing or heavy strap at the high-frequency end of the range.

The experiments also demonstrated something that the textbook equations for the rf resistance of an inductor imply: the Q of a coil, over wide limits, is more dependent on the size and shape of the coil than on conductor size.

Harry R. Hyder, W7IV
Scottsdale, Arizona

laser communications

Dear HR:

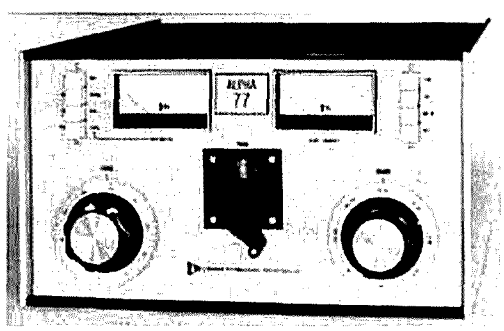
In regards to W4KAE's letter in the May issue, his claim to the first two-way laser QSO is a little late. There was a two-way laser QSO on February 25th, 1971 between WA8WEJ/Ø and W4UDS/Ø on a frequency of 475 THz using A3 modulation (*QST*, July, 1971, page 93). This precedes W4KAE's contact by almost 9 months.

Although W4KAE's two-way laser contact was not the first, it is definitely the DX record to date. Keep up the good work, Ralph.

Lee Yazell, WB9AIU
Pensacola, Florida

new products

alpha 77 linear amplifier



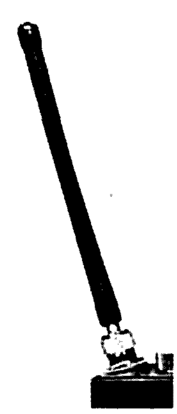
The Alpha 77 Linear power amplifier has several new design features which have been added since the unit was originally released. The new Alpha 77 now uses an air-cooled Eimac 8877/3CX1500A7 ceramic-metal triode with 4000 volts on the plate. This power tube has a conservative 1500 watts of plate dissipation and requires only 65 watts drive for full legal amateur power input.

Also new in the Alpha 77 is a grid excess-current relay which will kick out if the final tube is under-loaded or over-driven. This protects the tube and the input circuit. Primary power requirements for the amplifier are now 120/240 volts at 50/60 Hz, single phase, making the unit compatible with overseas power sources.

The Alpha 77 is designed and rated at 3000 watts PEP continuous commercial service and is available to amateurs who demand the ultimate in every respect; it loafs along at 2000 watts PEP. The Alpha 77 features a rugged bandswitch with 20-amp silver contacts, vacuum-variable tuning capacitor, silent vacuum relays, quiet forced-air cooling and metering in all circuits. The massive plate coil is silver soldered and heavily silver plated for efficiency. Husky toroid coils minimize coupling between pi-L network sections.

The Alpha 77 Power amplifier is built like a battleship with ¼-inch thick aluminum sides, and weighs in at 70 pounds. Modular assembly is used throughout so that the power supply, rf deck and control panel are easily removable. Second harmonic output is typically -50 dB, and the third-order intermodulation products are -35 dB below peak output. The Alpha 77 is manufactured by Ehrhorn Technological Operations, Inc., and distributed by Payne Radio, Box 525, Springfield, Tennessee 37172. For more information, use *check-off* on page 126.

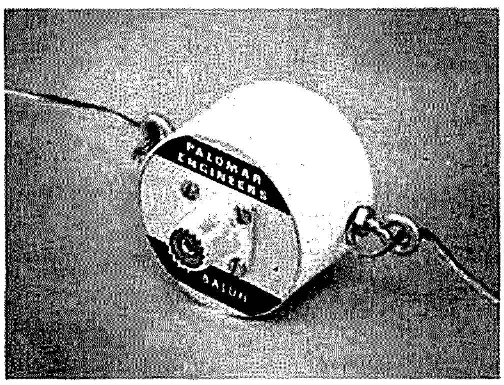
two-meter portable antenna



A new industrial type, continuously-loaded vhf portable antenna has been added to Antenna Specialists' high-performance amateur line. The new Antenna Specialists HM-5, designed to withstand the rough handling that makes telescopics impractical, is completely insulated and cannot accidentally be shorted out. It features a connector fitting that attaches directly to portable equipment with SO-

239 connectors. Power rating is 25 watts with nominal input impedance of 50 ohms. A companion model, the HM-4, is identical except for a standard 5/16-32 threaded-male mounting base. Both models are available from amateur distributors at a suggested ham net price of $5.95. For additional specifications, write to Amateur Department, The Antenna Specialists Company, 12435 Euclid Avenue, Cleveland, Ohio 44106 or use *check-off* on page 126.

palomar balun



Palomar Engineers has announced a new 1:1 Balun. It matches 50- or 75-ohm coaxial cable to center-fed dipole or inverted-vee antennas. By preventing radiation from the coax, a balun improves the antenna radiation pattern, reduces noise on receive, and helps prevent tvi, bci and rf feedback within the station.

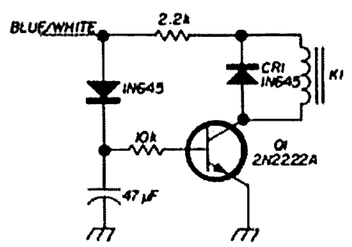
The Palomar Balun transformer is wound on a large ferrite toroid core and handles a full kilowatt from 1.7 to 30 MHz. The transformer is enclosed in a white plastic housing and is completely encapsulated to prevent moisture from entering. All hardware is stainless steel.

Eye bolts on the sides allow the balun to replace the center insulator of the antenna, and an eye bolt on top can be used to support the antenna. The balun is a compact 2¼ inch in diameter and 2 inches high. The unit is priced at $12.95 postpaid in the United States and Canada (plus 5% tax in California). A descriptive brochure is available. For more information, write to Palomar Engineers, Box 455, Escondido, California 92025 or use *check-off* on page 126.

short circuits

mobile touch-tone

In the article on mobile Touch-Tone operation in the August, 1972, issue, page 60, the capacitor and resistor across relay K1 can cause immediate failure of the transistor due to excessive current flow. When the transistor is first turned on, and the capacitor is not charged, current flow is about 500 mA since the 22-ohm resistor is the only thing that limits current. WB8NAT has suggested a modification which places the time-constant capacitor in the base circuit of the transistor. This has proved quite effective and according to several fm operators who have built it, the circuit works very well. The diode has been added to the transistor base circuit to prevent the dial from shorting out the charge on the time-constant circuit.



Improved time-constant circuit.

frequency-measuring oscillator

The frequency-measuring oscillator described in the April, 1972, issue was originally designed and built by Ben Christie, K2BF. A footnote should have been included in the article to that effect. It seems that Ben designed and built an fmo and sent it to a friend, Peter Petersen, K6MFS, in Long Beach, California. It was at K6MFS's house that author W6IEL saw the original fmo, realized it would make a good construction article and obtained full information from K2BF. The photographs in the article were taken by K6MFS, and are of the original fmo built by K2BF.

multimode i-f system

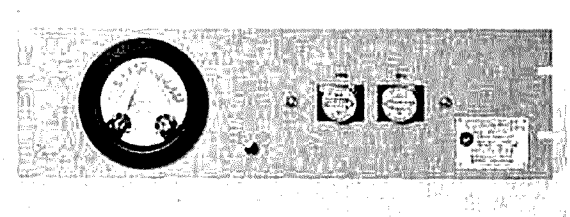
In the article on the multimode i-f system in the September, 1971, issue, several component designators were not included on the schematic. Capacitor C6 is the .001-μF capacitor from pin 14 to pin 15 of the NE560 PLL IC. C1 and C2 are the two 100-pF capacitors across L1 and L2. R1 is the 1k resistor from pin 2 to pin 3 of the MC 1596G IC. R3 is the 3.9k resistor connected to pin 9 of the MC1596G.

frequency scaler

In the circuit of the frequency scaler, fig. 2B, page 42 of the September, 1972 issue of *ham radio,* the .01-μF capacitor should be connected to both V_{cc} pins on the 95H90 IC. The circuit-connection dot was inadvertently left off by the draftsman.

direct-reading swr meters

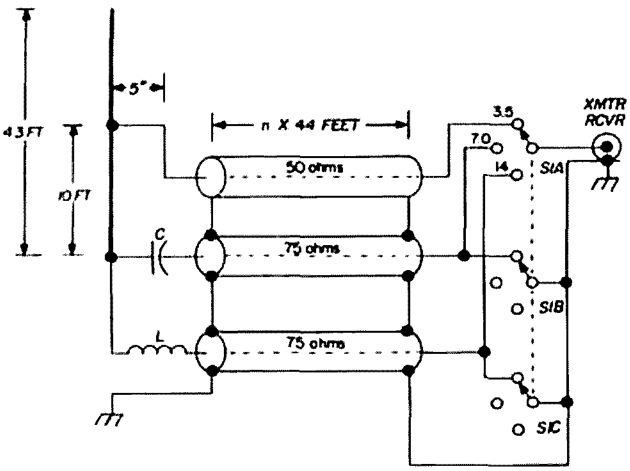
The meter scale for the direct-reading swr meter (fig. **1**, page 29, May, 1972) is incorrect. The author states that, "An swr of 1:1 is obtained if the reflected power is equal to zero." That would place the swr line labeled "1" directly between the zero on the *reflected* scae scale and the pivot of the reflected meter pointer. In fig. **1** that line is displaced to the left, so that a reflected reading *less* than zero would be required to obtain an swr of one. In addition, Thruline® is a registered trademark of Directional RF Wattmeters manufactured by Bird Electronic Corporation. Author WA4WDK's vswr meter was patterned after the Bird model 3122 Thruline® Wattmeter pictured here.



Bird Model 3122 Thruline wattmeter.

logic monitor

In the logic monitor circuit described in the April, 1972, issue (page 70) the output of the first section of the MC844 should be connected to the 1k pullup resistor.



LA1EI's three-band ground-plane antenna.

three-band ground plane

Fig. 6 of LA1EI's excellent article on the three-band ground plane in the May, 1972, issue is in error. The corrected drawing is shown above.

repeater control with simple timers

In **fig. 1** on page 47 of the September, 1972, issue, C1, the 400-pF capacitor, and the 33-ohm resistor should go to ground, not to Q of U1. Make the same change to **fig. 2** on page 48.

power supply ics

The RCA CA3055 voltage-regulator IC specified in the article on page 50 of the November, 1970 issue of *ham radio* has been replaced by the RCA CA3085. The CA3085 is a 12-mA device, while the CA3085A is a 100-mA device which is a plug-in replacement for the old CA3055. The CA3085B, a member of the same IC family, has the same ratings as the CA3085A, but can withstand higher input voltage surges. The price, incidentally, of the CA3085A is *less* than the CA3055, welcome news in view of the higher costs of nearly every commodity these days.

ham radio

index to volume V – 1972

This index covers all articles published in *ham radio* during 1972. The articles are listed alphabetically under each category along with the author, page number and month. Categories are: antennas and transmission lines; commercial equipment; construction techniques; fm and repeaters; integrated circuits; keying and control; measurements and test equipment; miscellaneous technical; power supplies; receivers and converters; RTTY; semiconductors; single sideband; transmitters and power amplifiers; vhf and microwave. Articles followed by (HN) appeared in the *ham notebook.*

antennas and transmission lines

audio

commercial equipment

construction techniques

fm and repeaters

integrated circuits

keying and control

measurements and test equipment

miscellaneous technical

power supplies

receivers and converters

RTTY

semiconductors

single sideband

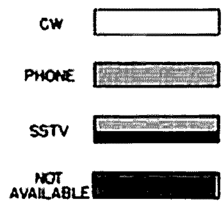
transmitters and power amplifiers

uhf and microwave

revised U.S. amateur

FREQUENCY ALLOCATIONS

effective November 22, 1972



from the editors of **ham radio** *and*

ROBOT

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Leading the way in Slow Scan Television